

TRANSIENT RESPONSE OF COMPENSATED
VIDEO INTERSTAGES

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A THESIS
Presented to
the Faculty of the Division of Graduate Studies
Georgia School of Technology

In Partial Fulfillment
of the Requirements for the Degree
Master of Science in Electrical Engineering

by
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August 1947

TRANSIENT RESPONSE OF COMPENSATED
VIDEO INTERSTAGES

Approved:

Date Approved by Chairman August 18, 1947

ACKNOWLEDGMENTS

I wish to express my sincerest thanks to Dr. W. A. Edson for his suggestion of the problem and also for his invaluable aid and patient guidance in its prosecution. I am indebted to Professor M. A. Honnell of the Department of Electrical Engineering and to Mr. James E. Garrett of the Photographic and Reproduction Laboratory at the State Engineering Experiment Station of Georgia School of Technology for their splendid cooperation.

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DEFINITIONS OF SYMBOLS USED

C_c	Coupling condenser.
C_g	Effective capacitance to ground of grid plus socket and wiring.
C_k	Cathode by-pass condenser.
C_p	Effective capacitance to ground of plate plus socket and wiring.
C_t	Total shunting capacitance of interstage.
e_1	Input signal voltage to stage.
e_2	Output voltage of stage.
f	Any frequency in the high end of the video band.
f_o	The "critical" frequency, a design parameter.
f_r	Resonant frequency.
f_w	Bandwidth
g_m	The dynamic transconductance.
G	Voltage gain at a frequency f .
G_1	Mid-frequency voltage gain.
K	A design parameter.
m	A design parameter.
r_p	The dynamic plate resistance.
R_c	Resistance of series peaking coil.
R_d	Damping resistor in parallel with an inductance.
R_g	Grid leak resistance.
R_k	Cathode resistance.
R_L	Plate load resistance.

R_s	Damping resistor in series with an inductance.
R_t	Parallel combination of r_p , R_L , and R_g .
Z	Impedance of a two-terminal coupling network at a frequency f .
Z_t	Parallel combination of r_p and Z .
ω	$2\pi f$.
Φ	Phase delay in radians caused by reactance in plate load circuit.
t	Time delay in seconds ($t = \Phi/\omega$).
Δt	Departure from constant time delay in seconds.

The meanings of symbols not defined here are clearly indicated on the diagrams to which they refer.

TRANSIENT RESPONSE OF COMPENSATED

VIDEO INTERSTAGES

INTRODUCTION

General Considerations

The usual function of an amplifier is to magnify the signals applied to its input without altering their waveforms in any way.*

This is no problem in the case of sine waves since they cannot be distorted by a linear device. A resistance-coupled voltage amplifier in which the tube is operated linearly is such a device. All frequencies are not amplified equally, but this cannot produce distortion if only a single frequency is present.

The amplification of non-sinusoidal waves is a different proposition since these consist of component frequencies which must be amplified equally and with unaltered relative phase angles if the waveforms are to be preserved.

The resistance-coupled circuit is the basic one which is used to amplify complex waves because it may easily be modified to make its response uniform over a wide range of frequencies.

*Special wave-shaping circuits in which non-linear operation is utilized to produce non-sinusoidal waves are an exception which will not be considered here.

The gain reduction at high frequencies, a major factor in the production of distortion of complex waves, is due to the presence of stray capacitances in shunt with the load circuit of the amplifier tube. These capacitances consist of the interelectrode capacitance of the tube and socket, plus the shunt capacitance in the wiring of the interstage network and the input capacitance of the next tube and its socket. The strays may be minimized by careful construction and a judicious choice of tubes, but they can never be completely eliminated.

A video* amplifier consists of the basic resistance-coupled circuit, compensated to produce uniform response over a very wide band of frequencies. Its function is the amplification of pulse signals, often of a transient nature, and the wide-band characteristics are necessary to insure uniform amplification of as many of the frequency components present in these signals as possible.** The high-frequency response, which is of prime importance, is usually improved by the addition of inductance to the coupling network in an attempt to cancel the effects of the stray shunt capacitance.

The problem of video amplifier design has assumed a

*These amplifiers were first developed for television applications, hence the term "video" which is derived from the Latin "videre" (to see).

**It is well known that a pulse may be resolved, by Fourier methods, into a great number of sinusoidal components, a sharp pulse having infinitely many.

new and greater importance since the advent of radar and numerous other electronic devices in which pulse amplification is essential.

Summary of Video Amplifier Requirements

It has been pointed out that an amplifier, in order to faithfully reproduce a pulse, would have to be capable of preserving the relative amplitudes and phase angles of all of its frequency components. These are the ideal characteristics, expressed in "steady-state" terms.

The ideal amplifier would have an infinite bandwidth, which is not physically realizable. It is possible, however, to secure acceptable performance with practical amplifiers having finite bandwidths. The minimum bandwidth required for acceptable performance depends upon the application for which the circuit is designed. It may be shown that the required bandwidth increases as the duration of the pulse to be amplified decreases.¹

A video amplifier must exhibit both a constant time delay and a uniform amplitude response. The amplitude response is generally considered uniform over a band if the gain variation is not more than ± 1 db. (± 10 per cent variation in amplitude). The time delay, t , is defined by

¹George B. Hoadley and William A. Lynch, "Transients in Coupling Circuits," Communications, 23, June, 1943 and July, 1943; Ernst A. Guillemin, Communication Networks, Vol. II (John Wiley & Sons, 1935), p. 486.

$t = \Phi/2\pi f$, where Φ is the phase angle of the amplifier at the frequency f . If Φ is expressed in radians and f in cycles per second, t is given in seconds. It is evident that if t is to be the same at all frequencies, the necessary condition for the preservation of the phase angles between the Fourier components of a pulse, Φ must be proportional to frequency. The actual value of the time delay is not important, the only requirement being that it be invariant with frequency.

As an example of video amplifier requirements in television applications, Zworykin and Morton² state that the transmission of a 441 line picture with an aspect ratio of 4 x 3 at 30 frames per second requires a bandwidth of at least 3 megacycles. However, a 4.5 megacycle bandwidth with the low-frequency limit below frame frequency (30 cycles per second) is preferred.

The requirements which must be met in other applications of video amplifiers are often even more exacting. Amplifiers with 10 megacycle bandwidths (gain constant to within ± 1 db.) have been developed.³

The Compensation Problem

The bandwidth limitations of resistance-coupled

²V. K. Zworykin and G. A. Morton, Television (John Wiley & Sons, 1940), p. 397.

³Ellsworth D. Cook, "A Wide-Band Oscilloscope," Proc. IRE, 31:410ff., August, 1943.

stages are well known⁴ and numerous schemes have been devised to compensate for poor response at both the high and low ends of the band. High frequency and low-frequency compensation are essentially separate problems. However the former is usually considered first in design because certain constants fixed by high-frequency requirements also tend to determine the low-frequency response.⁵

The choice of tubes is important in the design of a compensated interstage and it is found that pentodes are, in general, the most satisfactory in view of their high ratio of transconductance to interelectrode capacitance.⁶ A high transconductance is necessary to obtain sufficient gain and low interelectrode capacitances help to minimize the stray shunt capacitance, the factor which limits the high-frequency response of the amplifier.

According to Seeley and Kimball,⁷ low-frequency compensation is required below about 200 cycles and high-frequency compensation becomes necessary at frequencies

⁴Donald G. Fink, Principles of Television Engineering (McGraw-Hill, 1940), pp 210-219.

⁵D. L. Jaffe, "Wide-Band Amplifiers and Frequency Multiplication," Electronics, 15:56ff., April, 1942.

⁶Fink, op. cit., p. 238; Zworykin and Morton, op. cit., p. 403.

⁷Stuart W. Seeley and Charles N. Kimball, "Analysis and Design of Video Amplifiers," Part II, R.C.A. Review, 3:290ff., January, 1939.

above 100 kilocycles. This statement is not to be considered a hard and fast rule, the frequency limits mentioned being intended only to bring the reader into focus on orders of magnitude.

High-frequency compensation presents more difficulty, in general, than low-frequency compensation and consequently receives more attention in the literature. This is probably due, in part, to the fact that any high-frequency compensation system must be designed to combat the effects of stray shunt capacitance, a factor which can be minimized, but never completely eliminated. The low-frequency limitation is imposed by circuit elements which are actually selected by the designer, and are therefore better under his control. It also happens to be true that the important pulse transmission characteristics of an interstage are largely governed by its high-frequency response. The term "compensated interstage" is generally understood to mean "high-frequency compensated interstage" unless otherwise specified.

Seeley and Kimball⁸ point out that the compensation problem is complicated by the fact that an interstage which is designed to produce a constant time delay throughout the video band generally exhibits a non-uniform gain characteristic. They state in a more recent article⁹ that a network

⁸Stuart W. Seeley and Charles N. Kimball, "Analysis and Design of Video Amplifiers," R.C.A.Review, 2:171ff., October, 1937.

⁹Same as footnote 7.

which is effectively corrected for flat high-frequency amplitude response usually produces a tolerably uniform time delay and added time delay correction expedients are, therefore, unnecessary. This statement is not intended as a general rule, but merely as an approximation leading to performance which is satisfactory in many cases but unsatisfactory in others, depending upon the requirements. If a large number of identical stages are in cascade it is likely that a careful compromise between amplitude and phase response correction measures will have to be made. The overall gain characteristic is the product of those of the individual stages and the total time delay is the sum of the individual stage time delays, therefore each stage must be compensated as well as possible.

The Transient Point of View

The discussion, thus far, has outlined the video amplification problem in terms of amplitude and phase response, the so-called "steady-state" characteristics. Measurements of these quantities may be performed, using sine wave signals of constant amplitude, and the information gained will be of considerable value in predicting the amplifier's transient performance. Most interstage designs, particularly the more complicated ones, are based on these criteria.

High-frequency compensation systems generally involve the addition of inductances to the circuit. These tend to resonate with the capacitances present and oscillations

often result when transient signals are applied. Such troubles may be expected to reach distressing proportions in complex networks. A steady-state analysis does not account for these effects.

Furthermore, the steady-state point of view is not adequate for a consideration of the network response to signals which are not repetitive. Most television picture signals belong in this category. Evidently a more advanced type of analysis is required, specifically a transient solution of the amplifier's response.

As Fink¹⁰ puts it, "If the amplifier response to a generalized transient signal is determined, the performance of the amplifier for video amplification is more fully known than if reliance is placed simply on steady-state responses."

The commonly used generalized transient signal is the "Heaviside unit voltage" which is zero until a time T when it rises to unit amplitude instantly and remains there indefinitely. Uniform amplitude and phase responses from zero to infinite frequency are required for the completely undistorted passage of such a pulse.

Theoretical transient analyses are to be found in the literature,¹¹ particularly for the simpler networks, and

¹⁰Fink, op. cit., p. 245

¹¹Ibid., pp. 245-250 for a good summary. Guillemin, op. cit., pp. 461-506; N. W. McLachlan, "Reproduction of

will not be given here. They are quite tedious and the experimental approach is favored.¹²

Bedford and Fredendall¹³ present a method of transient response analysis which is simpler than the approaches of the other authors. A recurring square wave of suitable period is used as the applied voltage rather than the unit pulse, permitting the use of a Fourier series rather than the more formidable Fourier integral. The distorted output waveform of the network being studied is approximated, not by a series of harmonic sine waves, but by a series of displaced square waves, all having the original period.

The use of the square wave may be justified by the fact that the leading edges of both the square wave and unit pulse are distorted in essentially the same way by the amplifier. That is, the output pulse has a non-zero rise time or rounded corner. The sharp fall at the end of the square wave is distorted in the same way as the initial rise. This type of distortion is due almost entirely to non-uniform high-frequency gain and phase responses. The period of the square wave used should be just long enough

Transients by Television Amplifiers," Wireless Engineer, 13: 519ff., October, 1936; Eric A. Walker, "Transient Amplifier Analysis," Electronics, 12:39ff, November, 1939.

¹²Hoadley and Lynch, op. cit., for some experimental results.

¹³A. V. Bedford and G. L. Fredendall, "Transient Response of Multistage Video-Frequency Amplifiers," Proc. IRE, 27:277ff., April, 1939.

to insure that the principal transient effects will be over before the polarity reversal at the end of a half-period takes place. This provision sets the lower limit on the period.

The analysis is carried out over the time interval during which the output pulse is changing in value from zero to a steady amplitude equal to its final value.

Low-frequency distortion of the unit pulse is evidenced by a gradual drop in amplitude a considerable time after the initial rise. This is eliminated from the analysis by limiting the period of the square wave so that the low-frequency effects do not have time to occur. Low-frequency analysis can be carried out separately if desired.

It will be shown that the use of a square wave to approximate a unit pulse is most advantageous in the laboratory as well as in a mathematical treatment.

The Scope of the Thesis

It is the purpose of this thesis to present the results of square wave tests on most of the high-frequency compensated interstage networks which appear in the literature.

To simplify the experimental work it was decided to start with an amplifier capable of faithful reproduction of the square wave used. Distortion was deliberately introduced by the addition of accurately known amounts of shunt capacitance. The use of relatively large shunt capacitances

made it possible to use signal frequencies low enough to eliminate the effects of all actual stray capacitances existing in the amplifier. Thus all factors involved in the problem were accurately controlled.

A brief outline of the train of thought followed in designing the networks is first presented and the values of circuit constants recommended by several authors are summarized. An illustration of the interpretation of square wave test results is given in terms of steady-state characteristics.

The experimental work is then described and the results are presented in the form of oscillograms. Finally the results are reviewed and conclusions are drawn regarding the relative merits of the networks tested.

From the oscillograms it is possible to select a network to meet a given set of engineering requirements. Simple means for obtaining the numerical values of the required circuit elements are presented.

INTERSTAGE DESIGNS FROM THE LITERATURE

Background

It has been common practice to base interstage designs on steady-state criteria, so the information presented in this section will be expressed in those terms. The following statements by Sarbacher and Edson¹⁴ will direct subsequent thinking along the proper channels: "It should be emphasized that the only practical limitation on the upper frequency which may be amplified by a video amplifier is set by the shunting capacitances. If it were somehow possible either to remove or to annul these capacitances it would be possible to obtain a very large amplification per stage and to extend the amplification to frequencies at which lead inductance or transit time effects become important."

In this same connection, Wheeler¹⁵ has derived the relation that the maximum uniform gain obtainable with one tube is

$$G_{\max} = \frac{g_m}{\pi f_w \sqrt{C_p C_g}}$$

¹⁴Robert I. Sarbacher and William A. Edson, Hyper and Ultra-High Frequency Engineering (John Wiley & Sons, 1943), p. 478.

¹⁵Harold A. Wheeler, "Wide-Band Amplifiers for Television," Proc. IRE, 27:429ff., July, 1939.

where f_w is the band-width.

Consider the resistance-coupled circuit shown in Figure 1(a). At high frequencies the equivalent circuit of Figure 1(b) represents the prevailing conditions. All networks described will be drawn in the form of high-frequency equivalent circuits.

Referring to Figure 1(b),¹⁶ the output voltage, e_2 , is given by

$$e_2 = e_1 g_m Z_t = e_1 g_m \frac{\frac{R_t}{j\omega C_t}}{R_t + \frac{1}{j\omega C_t}}$$

or

$$e_2 = e_1 g_m \frac{R_t - j\omega C_t R_t^2}{1 + \omega^2 C_t^2 R_t^2}.$$

The high-frequency gain, G , is then

$$G = \frac{R_t g_m}{\sqrt{1 + \omega^2 C_t^2 R_t^2}}$$

with a phase angle $\Phi = -\tan^{-1}(\omega C_t R_t)$.

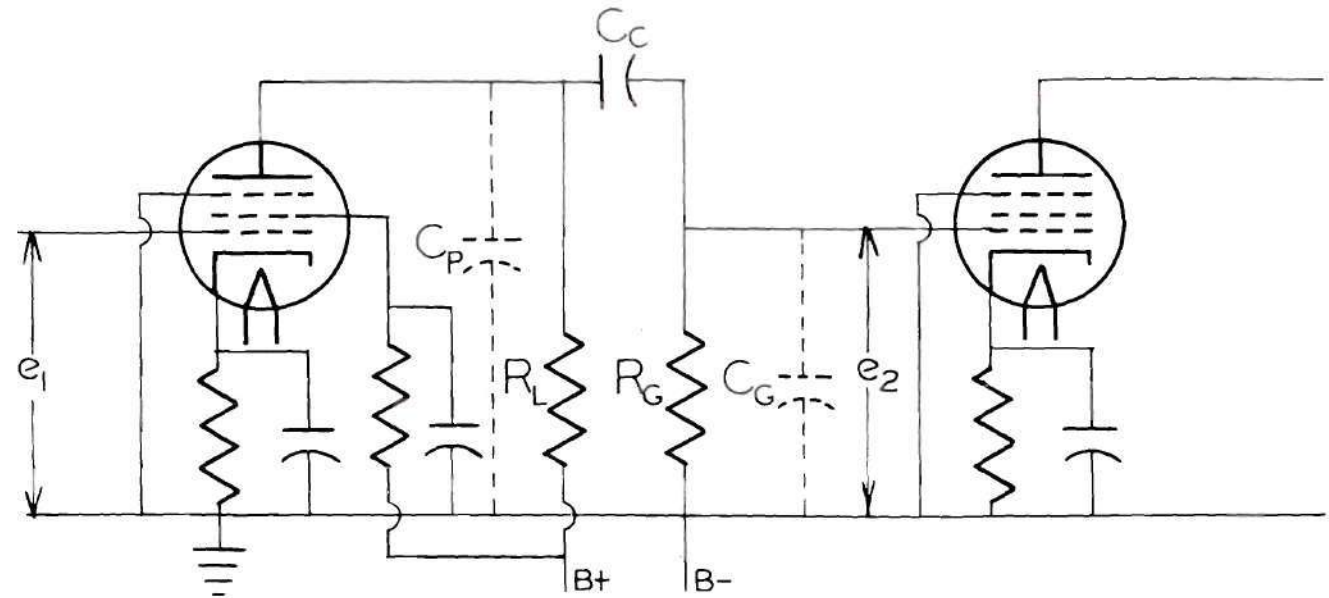
It is convenient to define a "critical frequency", f_o , by

$$f_o = \frac{1}{2\pi C_t R_t}.$$

At this frequency the reactance of C_t is equal in magnitude

¹⁶Zworykin and Morton, op. cit., pp. 398-402.

(a) Schematic diagram



(b) High-frequency equivalent circuit

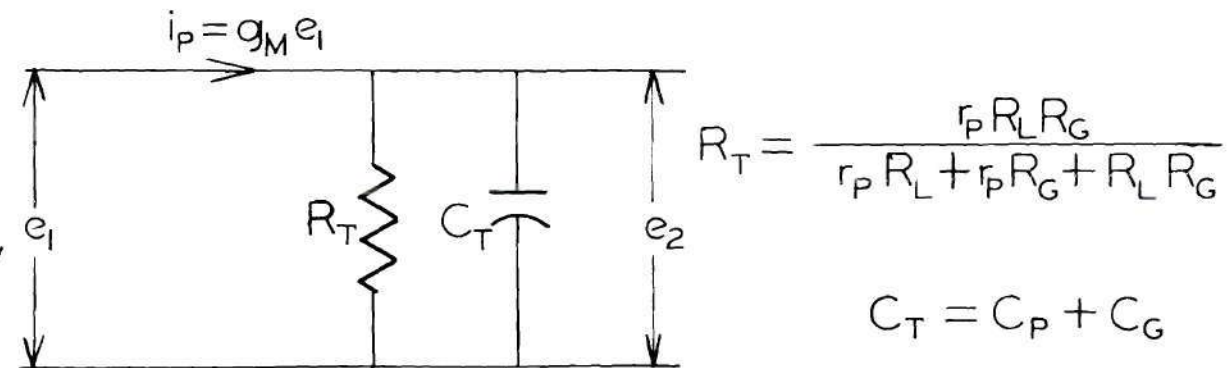


FIG. 1.— UNCOMPENSATED RESISTANCE—COUPLED STAGE

to R_t . The mid-frequency gain, G_1 , is given by $G_1 = g_m R_t$ since C_t is neglected at these frequencies. It is easily verified, by substitution, that the gain will be $0.707G_1$ at the frequency f_0 .

It may also be shown that the amplifier phase angle, Φ , is 45 degrees at the frequency f_0 . Since time delay is given by $t = \Phi/2\pi f$, the time delay at f_0 is $0.125/f_0$. This value would hold for any frequency in an amplifier having uniform phase characteristics. In the uncompensated stage, however, the time delay decreases with decreases of frequency so that a time delay error exists. Seeley and Kimball¹⁷ state that the time delay error Δt is approximately equal to $0.034/f_0$ at the frequency f_0 . Zworykin and Morton¹⁸ say that at a frequency $f = 0.48f_0$ the gain is 90 per cent of G_1 and the time delay error is $0.005/f_0$.

It is evident that $G = g_m Z_t$. In a pentode r_p is very large so that Z_t is approximately equal to the impedance, Z , of the coupling network. Therefore, $G = g_m Z$ and Φ is essentially the angle of Z .

If the frequency range is to be extended, without adding any elements to the circuit, f_0 must be increased. C_t is fixed, so the only recourse is to decrease R_t . This must be done by reducing R_L because r_p is fixed and a

¹⁷Same as footnote 7.

¹⁸Zworykin and Morton, op. cit., p. 409.

reduction in R_g will degrade the low-frequency response.¹⁹ A reduction in R_L produces a proportional decrease in gain so the method has serious limitations.

It is obvious that the simple resistance-coupled interstage is not adequate for video applications. More complex networks must be considered. In all cases R_g will be very large compared to R_L so its effect will be neglected.

It is important to distinguish between two-terminal and four-terminal interstages. The former is a network in which C_p and C_g are lumped. In the latter, C_p and C_g are at opposite ends of the network; they are separated by compensating elements, a single inductance in the simplest case.

The problem of maintaining a constant gain over a wide band of frequencies with a two-terminal network resolves itself into designing a coupling network such that Z does not vary with frequency over the band. To maintain zero time delay error the angle of Z must be directly proportional to frequency.

There are two general points of view concerning the design of compensation networks. They will be treated in detail later, but may be mentioned briefly here. The first method consists of selecting a logical network configuration from previous experience. For example, to compensate for an impedance which decreases as the frequency increases it

¹⁹Fink, op. cit., pp. 216-217.

is reasonable to insert an inductance, which exhibits the opposite behavior. The desired interstage characteristics are then obtained by imposing suitable conditions on mathematical expressions which relate the governing factors to the performance. Solutions of the resulting equations yield the required design information.

A more recent and generalized approach involves the application of two-terminal and four-terminal filter theory. Wheeler²⁰ has done considerable work on this aspect of the problem. A coupling network designed by this method consists of a low-pass filter terminated in R_L . The shunt capacitances C_p and C_g are included in the network. The filter method is of considerable advantage, particularly in four-terminal designs. The objectives sought are the same, regardless of the method. In the two-terminal case the idea is to make the magnitude of the self-impedance, Z , of the network constant over the video band and the angle of Z directly proportional to frequency. In the four-terminal case these same properties should be exhibited by the network transfer impedance.

The filter approach facilitates the design of complex interstages which are beyond the scope of the other method. Improved performance is to be expected as more elements are added but this process cannot be carried on

²⁰Wheeler, op. cit.

indefinitely.²¹

Two-Terminal Interstages

Network 1., a simple compensated two-terminal interstage is illustrated in Figure 2(a). As indicated on the diagram, this system is referred to as "shunt compensation." The analysis will be outlined rather briefly.²² The impedance, Z , is given by

$$Z = \frac{\frac{R_L}{2\pi f C_t} + j\left(\frac{L}{C_t} - 4\pi^2 f^2 L^2 - R_L^2\right)}{2\pi f C_t \left[R_L^2 + \left(2\pi f L - \frac{1}{2\pi f C_t}\right)^2 \right]} \quad (1)$$

The gain may be written as

$$G = g_m Z = g_m \frac{\sqrt{\frac{R_L^2}{4\pi^2 f^2 C_t^2} + \left(\frac{L}{C_t} - 4\pi^2 f^2 L^2 - R_L^2\right)^2}}{2\pi f C_t \left[R_L^2 + \left(2\pi f L - \frac{1}{2\pi f C_t}\right)^2 \right]} \quad (2)$$

with an angle

$$\phi = \tan^{-1} \frac{\frac{L}{C_t} - 4\pi^2 f^2 L^2 - R_L^2}{R_L / 2\pi f C_t} \quad (3)$$

²¹Ibid

²²See E. W. Herold, "High-Frequency Correction in Resistance-Coupled Amplifiers," Communications, 18:11ff., August, 1938. Also Zworykin and Morton, op. cit., pp. 405-409. Equations and numerical results were taken from the latter.

For ease of manipulation, substitute $f_o = 1/2\pi C_t R_L$,
 $K = L/C_t R_L^2$ and $G_1 = g_m R_L$. G_1 is the stage gain at moderately low frequencies where the effect of C_t is negligible. It is often referred to as the "mid-frequency" gain. The gain as a function of frequency becomes

$$\frac{G}{G_1} = \sqrt{\frac{K^2(f/f_o)^2 + 1}{K^2(f/f_o)^4 - (2K - 1)(f/f_o)^2 + 1}} \quad (4)$$

and

$$\phi = \tan^{-1} \frac{f}{f_o} \left[(K - 1) - K^2 \frac{f^2}{f_o^2} \right] \quad (5)$$

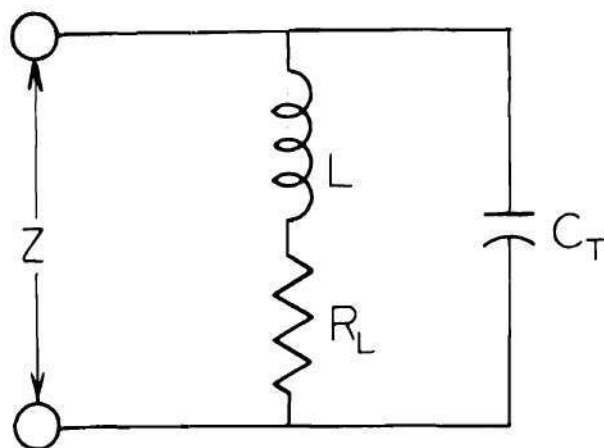
If $K = 1/2$ when $f = f_o$, Equation (4) gives $G/G_1 = 1$. This value of K gives 100 per cent amplitude response at f_o but the gain ratio is greater than unity at frequencies slightly below f_o and then falls rapidly. The time delay is not constant either. A more satisfactory value for K may be determined as follows:

The square of the network impedance is expressed as a rational fraction in which numerator and denominator are power series of the frequency:

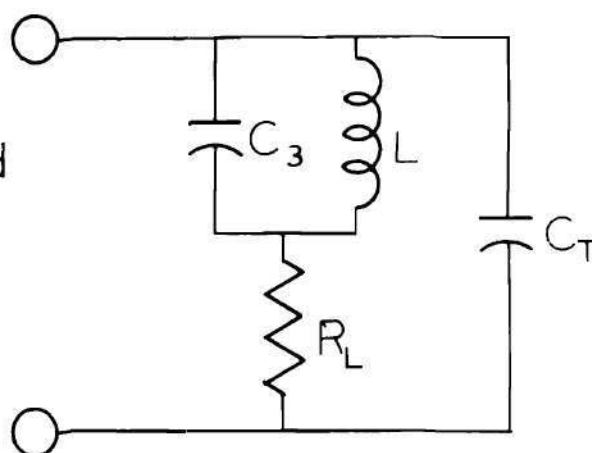
$$\left| Z(f) \right|^2 = \left| Z(0) \right|^2 \frac{1 + a_1 f^2 + a_2 f^4 + \dots}{1 + b_1 f^2 + b_2 f^4 + \dots} \quad (6)$$

If $a_1 = b_1$, $a_2 = b_2$, etc., the response does not vary with

(a) Network 1.— Simple
"shunt compensation."



(b) Network 4.—Modified
shunt compensation.



Note: Networks 2 and 3 are described in the text.

(c) Network 5.

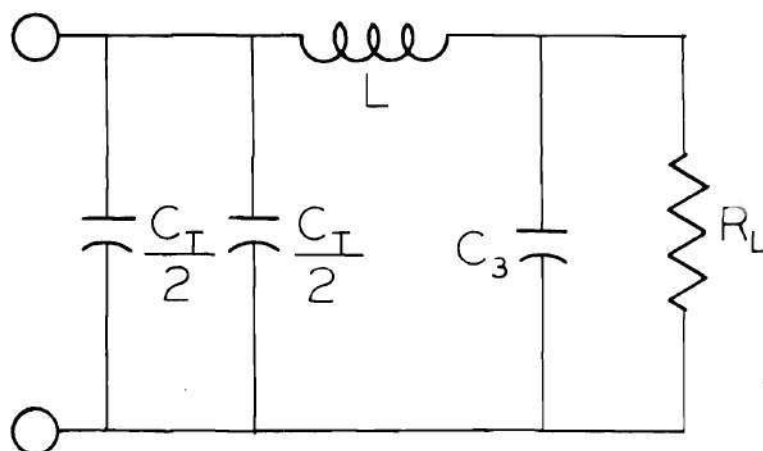


FIG. 2— THREE BASIC TWO-TERMINAL
INTERSTAGE NETWORKS

frequency. Similarly, $d\phi/df$ may be expressed as a ratio of two polynomials and made constant by making coefficients of like terms in numerator and denominator equal.

In effect, Equation (4) has been reduced to the form of (6). Since the denominator contains $(f/f_0)^4$ while the numerator does not, this type of coupling can never give ideal performance. However, if $(f/f_0)^4$ can be neglected in comparison to $(f/f_0)^2$ the response will be constant if $K^2 = 1 - 2K$ or $K = 0.414$. This approximation is acceptable over a considerable range of frequencies. If $K = 0.414$, the relative gain at $f = f_0$ is theoretically 92 per cent.

If Equation (5) is differentiated with respect to frequency and coefficients of like terms are equated, constant delay is obtained when $K^3 + 3K - 1 = 0$ or $K = 0.32$. Obviously a compromise value of K must be used. If $K = 0.37$, the response at f_0 is 90 per cent G_1 and $\Delta t = 0.003/f_0$ up to f_0 .

Other authors give similar information,²³ sometimes expressed in different terms. Several seem to favor conditions corresponding to $K = 0.414$, particularly from constant time delay considerations, although the reduction of K from 0.5 to 0.414 produces about 15 per cent less gain in the

²³ Seeley and Kimball (see footnote 7); William H. Freeman, "Shunt-Peaking Compensation," Electronics, 13:35ff., January, 1940; Alexander B. Bereskin, "Improved High-Frequency Compensation for Wide-Band Amplifiers," Proc. IRE, 32:608ff., October, 1944.

region of f_0 .

Network 2. is not illustrated, but consists of ordinary resistance-coupling with a rather small condenser, C_K , placed in shunt with the cathode resistor R_K .*

At low frequencies the parallel impedance of C_K and R_K is essentially R_K , but at high frequencies it is negligible. Therefore, negative current feedback²⁴ which decreases as the frequency rises is employed. The voltage applied to the coupling network is larger at the higher frequencies and this tends to compensate for the effect of C_t . The value of C_K was tentatively chosen so that the time constant $R_K C_K$ would be equal to $R_L C_t$.

Network 3.²⁵ is a combination of Networks 1 and 2, that is it consists of Network 2 with a compensating inductance added in series with R_L .

Network 4., Figure 2(b), is a form of shunt compensation in which the performance is improved by the use of a parallel LC combination instead of a simple inductance. The object is to produce the effect of an inductance whose value increases with increase of frequency. The LC combination behaves in this manner at frequencies below its natural re-

* Referring to Figure 10, R_K is the 1000 ohm resistor in the cathode circuit of the 6SJ7.

²⁴ Zworykin and Morton, op. cit., pp. 412-413.

²⁵ McLachlan, op. cit.

sonance. Such behavior is desirable because a larger value of compensating inductance is needed at frequencies above f_0 than at the lower frequencies.²⁶

Design recommendations:

(a) Terman²⁷

Relative impedances at f_0 :

$$R_L: 1.2 \quad C_t: 1.0 \quad L: 0.54 \quad C_3: 1.5$$

(b) Herold²⁸

$$2\pi f_0 C_t R_L = 1.25 \quad L = 0.38 C_t R_L^2 \quad C_3 = 0.3 C_t$$

Herold claims that these conditions produce a relative gain of 90 per cent at f_0 with a time delay variation, Δt , of $0.008/f_0$ up to that frequency.

(c) Bode²⁹

Values are given on a per unit basis (See Appendix I). They were derived from filter theory.

$$R_L = 1.0, \quad C_t/2 = 1.0, \quad C_3 = 1.335, \quad L = 0.75$$

The next four networks (5, 6, 7 and 8) are described

²⁶Sarbacher and Edson, op. cit., p. 481

²⁷Frederick E. Terman, Radio Engineers' Handbook (McGraw-Hill, 1943), p. 418.

²⁸Herold, op. cit.

²⁹Hendrik W. Bode, Network Analysis and Feedback Amplifier Design (D. Van Nostrand, 1945), p. 412.

by Bode.³⁰ The method of obtaining actual element values from the numbers given is explained in Appendix I. Filter terminology is used to some extent. The application of filter theory to interstage design will be outlined later and reference to pertinent material in the literature will be made at that time.

Network 5 is illustrated in Figure 2(c).

$R_L = 1.0$, $C_t/2 = 1.0$, $C_3 = 1.0$, $L = 1.80$ or 1.70

Network 6, Figure 3(a), is described by Bode as a conventional filter network in which matching to the load resistor, R_L , is accomplished by a single m-derived termination.

$R_L = 1.0$, $C_t/2 = 1.0$, $L = 1.60$ (1.50 by cut-and-try),

$L_2 = 1.067$, $C_3 = 0.600$

Terman³¹ and Jaffe³² give essentially the same values for this network.*

Network 7, Figure 3(b), is a filter configuration which was designed by "cut-and-try methods." Two sets of

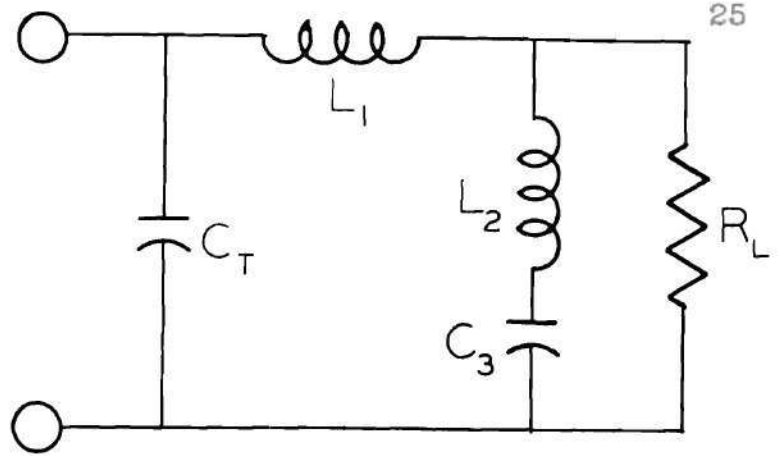
³⁰Loc. cit.

³¹Terman, Loc. cit.

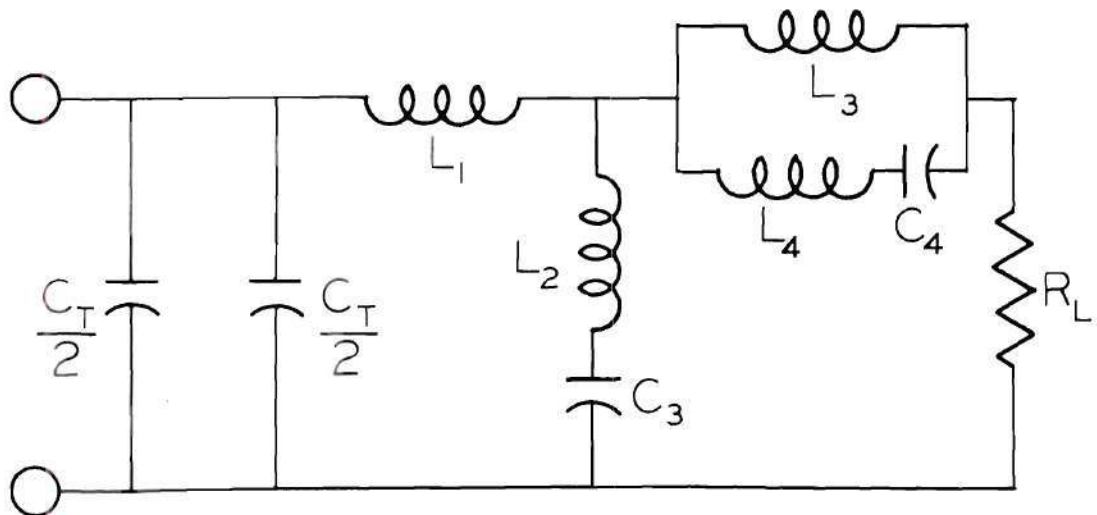
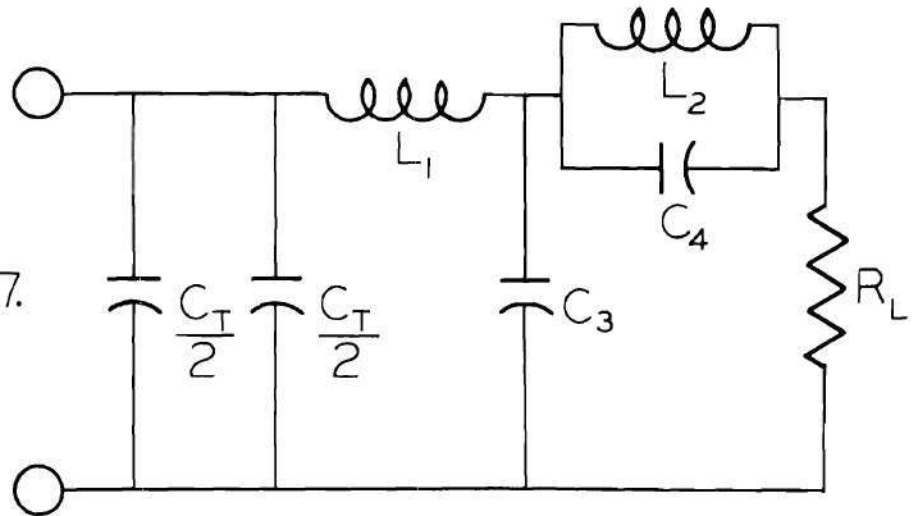
³²Jaffe, op. cit.

*The network was first encountered in Terman's handbook and the diagram was redrawn, in modified form, from his Figure 51(d). This explains why a single terminating capacitance, C_t , is shown rather than two of value $C_t/2$ each as in the other diagrams in this group.

(a) Network 6.



(b) Network 7.



(c) Network 8.

FIG. 3.—THREE TWO-TERMINAL NETWORKS OF INCREASED COMPLEXITY

values are given.

$$(a) \quad R_L = 1.0, \quad C_t/2 = 1.0, \quad L_1 = 2.000, \quad L_2 = 0.600, \\ C_3 = 1.600, \quad C_4 = 1.067$$

$$(b) \quad R_L = 1.0, \quad C_t/2 = 1.0, \quad L_1 = 1.9314, \quad L_2 = 0.600, \\ C_3 = 1.593, \quad C_4 = 0.8645$$

Network 8, Figure 3(c), is a filter network in which matching to R_L is obtained with a double m-derived termination.

$$R_L = 1.0, \quad C_t/2 = 1.0, \quad L_1 = 1.732, \quad L_2 = 0.4670, \\ L_3 = 0.2989, \quad L_4 = 0.3292, \quad C_3 = 1.022, \quad C_4 = 1.45$$

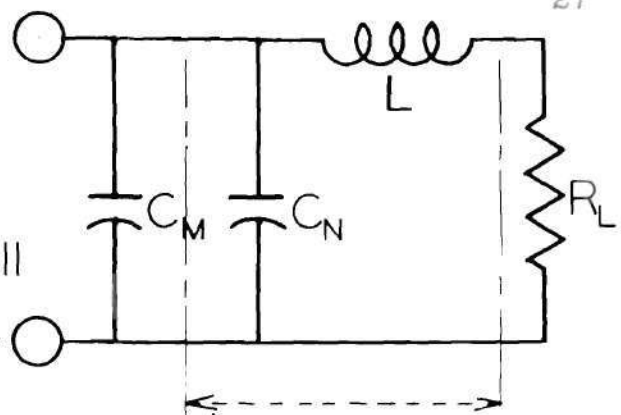
Two-Terminal Interstage Networks Considered as Filters³³

It has been pointed out that a two-terminal compensated interstage network must exhibit a self-impedance of constant magnitude and angle directly proportional to frequency over the band to be amplified. It is possible to approach such performance with specially designed low-pass filters. From another point of view, networks designed by other methods to meet the requirements may be considered as filters. The values of the circuit constants will be essentially the same regardless of the point of view adopted in executing the design.

³³Wheeler, op. cit.; Bode, op. cit., Chapter XVII; See Terman, op. cit., pp. 423-426 for a good summary. The diagrams of Figure 4 were redrawn from Terman's Figures 55 and 58(a). For a review of filter theory, consult T. E. Shea, Transmission Networks and Wave Filters (D. Van Nostrand, 1929), Chapter VII.

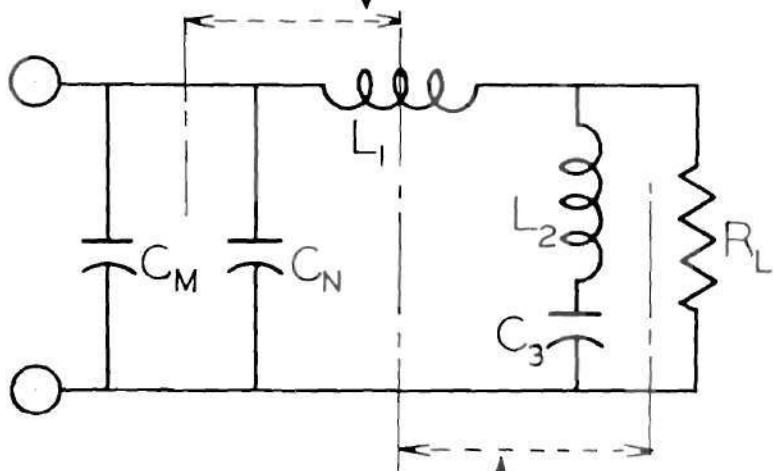
(a) Network 1.

Note: $C_T = C_M + C_N$ in all networks shown here.



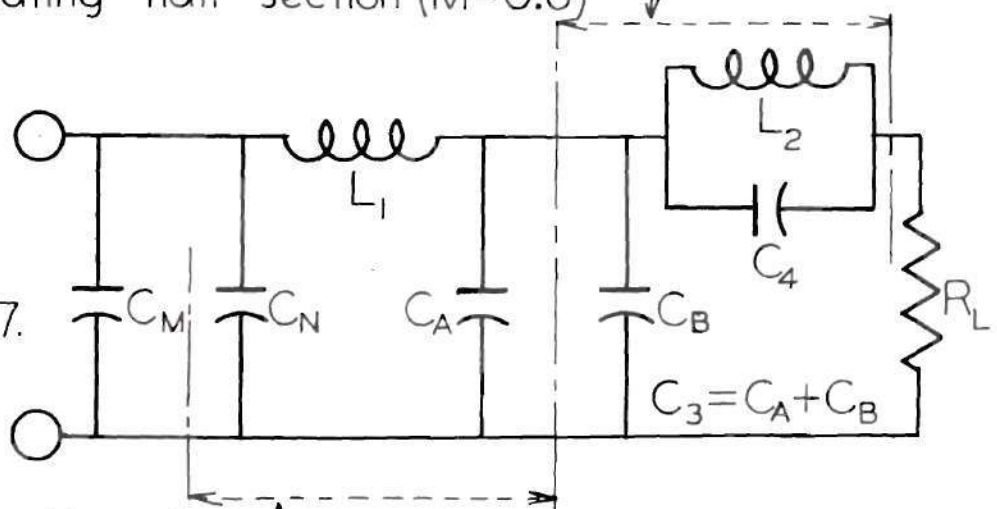
Constant K half-section

(b) Network 6.



Terminating half-section ($M=0.6$)

(c) Network 7.



Constant K section

FIG.4.- TWO-TERMINAL COUPLING NETWORKS SHOWN AS LOW-PASS FILTERS

The basic two-terminal configuration used in compensated interstages consists of a modified constant- k π section which is matched, on an image impedance basis, by an m -derived half section ($m = 0.6$), to a load consisting of R_L . Three of the networks which have been described are redrawn as filters in Figure 4. It is seen that the terminating half section is omitted in Network 1 and a constant- k half section, rather than a full section, is used in Network 6.

The modification to the constant- k section consists of shunting an auxiliary capacitance, C_M , across the input terminals. The total shunting capacitance across these terminals will be the same as the full shunt capacitance of a constant- k section if C_M is equal to C_N , the normal filter input capacitance. Terman shows that if the filter were perfectly matched to R_L , this value of C_M would give an impedance which is precisely constant to the cut-off frequency of the filter and a fairly constant time delay. Compromise values of C_M may be used to reduce the time delay variation.

It should be emphasized that capacitance is not actually added to the interstage, but that C_M consists of a portion of the capacitance, C_t , which is already present in the circuit.

Four-Terminal Interstages

Network 9, Figure 5(a), is the simplest four-terminal interstage, and the compensation which it provides is known

as "series peaking." The following analysis and numerical results are taken from Seeley and Kimball.³⁴

Referring to Figure 5(a), a constant-current generator is indicated, its output being $g_m e_g$. The voltage, e_1 , across R_L and C_p in parallel is given by

$$e_1 = \frac{g_m e_g R_L}{\sqrt{1 + (2\pi f C_p R_L)^2}}$$

This voltage is applied to L and C_g in series and the voltage drop across C_g tends to remain constant due to resonance effects in the series LC circuit.

The design procedure given by Seeley and Kimball will now be presented. First, let $C_g/C_p = m$. The m used here is not the parameter used in filter work. The capacitances should be adjusted so that $m \geq 2$, but it is not desirable to add capacitance to the network for this purpose since a loss of gain would result. The blocking condenser, C_c , which has appreciable capacitance to ground (Figure 1), may be placed at either end of L to facilitate the adjustment of m .

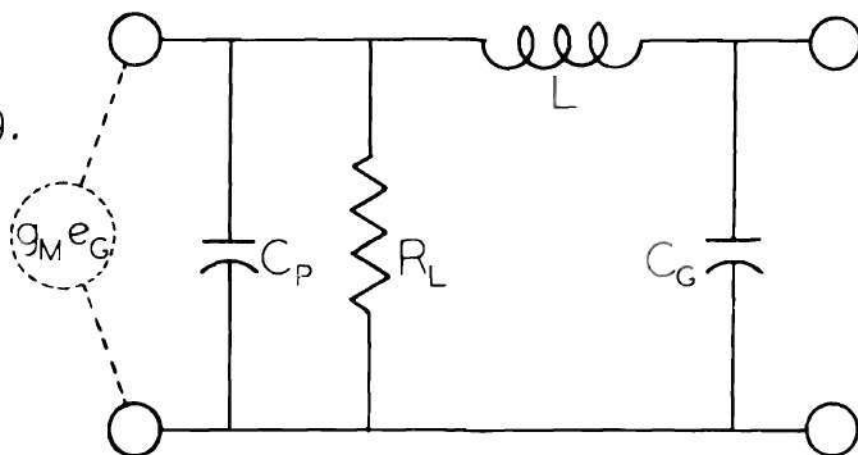
Let

$$f_r = \frac{1}{2\pi\sqrt{LC_p}}$$

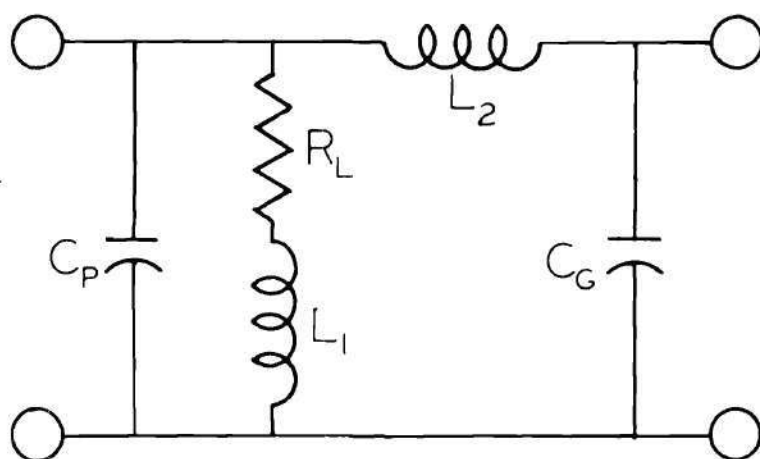
³⁴Same as footnote 7.

(a) Network 9.

"Series
peaking"



(b) Network 10.—
Combination of
series and shunt
peaking.



(c) Network 11.

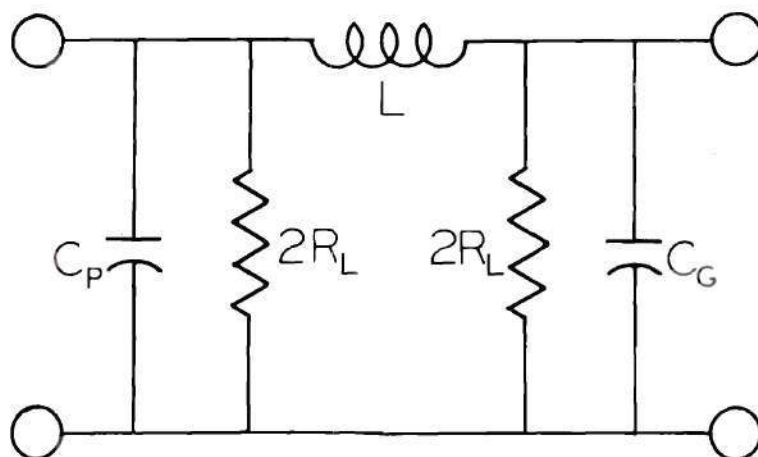


FIG. 5.— THREE BASIC FOUR-TERMINAL
INTERSTAGE NETWORKS

This is the resonant frequency of L and C_p . Choose L so that $f_r = \sqrt{2}f_o$, where f_o is defined as before by

$$f_o = \frac{1}{2\pi R_L C_t}$$

This gives

$$L = \frac{1}{2(2\pi f_o)^2 C_p}$$

With $m = 2$ and $f_o = f_r/\sqrt{2}$, let

$$R_L = 2\pi f_o L = \frac{1}{2(2\pi f_o) C_p}$$

Then

$$R_L = \frac{1}{2(2\pi f_o) \left(\frac{C_p + C_g}{3} \right)} = \frac{1.5}{2\pi f_o C_t}$$

Should $m = 1/2$, the network is reversed, that is, R_L is connected across the output terminals or across the smaller terminating capacitance. This condition does not occur in the usual amplifier.

In general, the design equations, when $f_o = \sqrt{2} f_r$ are:

$$R_L = \frac{1}{\sqrt{2m} 2\pi f_o C_p}, \quad L_2 = \frac{1}{2(2\pi f_o)^2 C_p}$$

It is stated by Seeley and Kimball that the following design will result in gain and time delay which are essentially constant out to the frequency f_0 :

$$C_g/C_p = 2, \quad R_L = \frac{1.5}{2\pi f_0 C_t}, \quad L = 0.67 C_t R_L^2$$

The Q of the coil should be greater than 20. They claim that the gain is 50 per cent greater in this circuit than in a shunt peaking circuit having the same C_t . The time delay is greater in a series peaking circuit than in a shunt peaking circuit, but this is of no consequence since it is more constant over the band. The maximum departure from constant time delay exhibited by the circuit just described is given as $\Delta t = 0.0113/f_0$ up to f_0 cycles.

Herold³⁵ gives the following design equations for this network:

$$2\pi f_0 C_t R_L = 1.23, \quad L = 0.56 C_t R_L^2, \quad C_p = 0.20 C_t, \quad C_g = 0.80 C_t$$

He says that the time delay variation, Δt , is $0.008/f_0$ and that the gain is 90 per cent of G_1 at the frequency f_0 .

Network 10, Figure 5(b), is a combination of series and shunt peaking. Seeley and Kimball³⁶ recommend the following design:

³⁵Herold, op. cit.

³⁶Same as footnote 7.

$$R_L = \frac{1.8}{2\pi f_0 C_t}, \quad L_1 = 0.12 C_t R_L^2, \quad L_2 = 0.52 C_t R_L^2, \quad C_g/C_p = 2$$

They claim that, due to an increase of about 80 per cent in the value of R_L , this circuit gives 80 per cent more gain than Network 9 gives over the same band. The Δt figure given is $0.015/f_0$ up to f_0 cycles.

Herold³⁷ gives practically the same design equations, the only difference being a very slight one in the capacitance ratio. He recommends $C_p = 0.34C_t$ and $C_g = 0.66C_t$. The performance claimed is 90 per cent relative gain at f_0 and a time delay variation of $0.009/f_0$.

The next six networks (11, 12, 13, 14, 15 and 16) were not found in the literature. They consist of modified forms of Networks 9 and 10.

Network 11, Figure 5(c), is a modified form of Network 9 in which two load resistors, each of value $2R_L$, are used, one at each end of the network. The total effective load resistance is, therefore, still R_L .

Network 12, Figure 6(a), consists of Network 10 with the same modification which was made to Network 9 to form Network 11.

Networks 13, 14, 15 and 16 are nothing more than Networks 9 and 10 with damping resistors added. The object is to find a method of increasing the circuit damping so

³⁷Herold, op. cit.

that large amounts of series peaking inductance could be used without producing oscillations in these networks.

Network 13 consists of Network 9 with a damping resistor, R_s , in series with L .

Network 14 consists of Network 9 with a damping resistor, R_d , in parallel with L .

Network 15 consists of Network 10 with a damping resistor, R_s , in series with L_2 .

Network 16 consists of Network 10 with a damping resistor, R_d , in parallel with L_2 .

Network 17, Figure 6(b) is a combination of Networks 4 and 9. Terman³⁸ gives the following values for the relative impedances at f_0 :

$$R_L: 1.5 \quad C_p: 1.25 \quad C_g: 1.0 \quad L_1: 0.675 \quad L_2: 2.25$$

$$C_3: 1.875$$

He claims that the gain of this circuit is 2.25 times as great as that of a shunt compensated stage (Network 1) with $K = 1/2$. It is also stated that the gain variation over the band is 4 per cent and that the maximum time delay error is $0.055/f_0$.

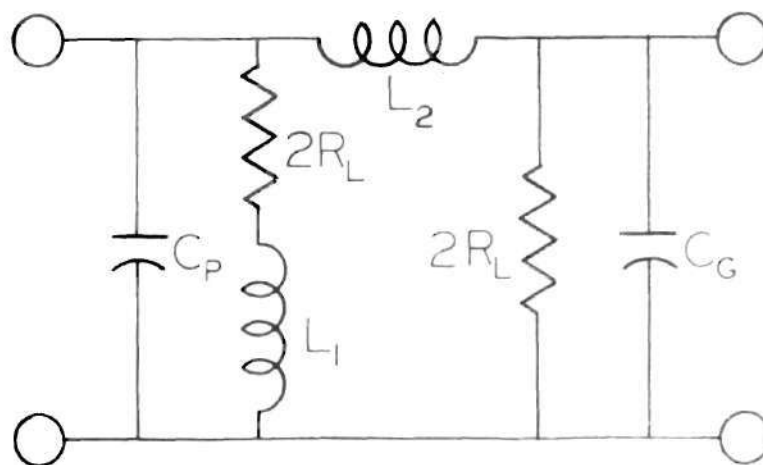
Bode³⁹ and Jaffe⁴⁰ give the same design values for

³⁸Terman, op. cit., p. 422.

³⁹Bode, op. cit., p. 430.

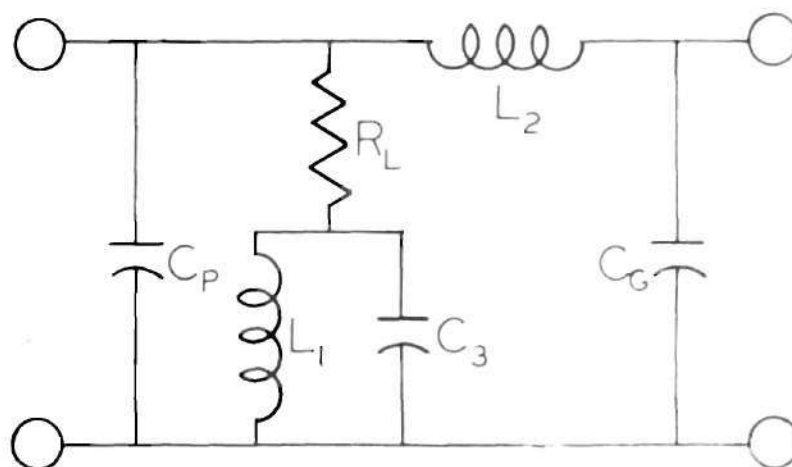
⁴⁰Jaffe, op. cit.

(a) Network 12.



Note: Networks 13, 14, 15 and 16 are described in the text.

(b) Network 17.



(c) Network 18.

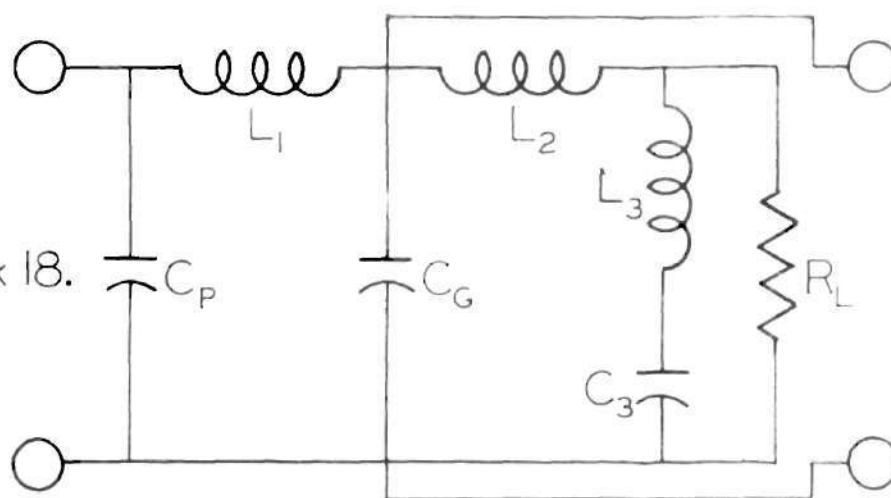


FIG. 6.— THREE FOUR-TERMINAL NETWORKS OF INCREASED COMPLEXITY

this network. The former points out that, in this circuit, a mid-shunt derived half section is used to match the resistance R_L .

Network 18, Figure 6(c), is presented by Terman⁴¹ with the following relative impedances at f_0 :

$R_L: 1.5 \quad C_p: 1.0 \quad C_g: 1.0 \quad L_1: 2.25 \quad L_2: 1.8 \quad L_3: 1.2$

$C_3: 3.33$

Bode⁴² makes the same recommendations and describes the network as a filter structure in which the mid-shunt image impedance presented by the input terminals of the driven tube (grid to cathode) is matched to R_L by a mid-series derived half section with $m = 0.6$.

Four-Terminal Interstage Networks Considered as Filters⁴³

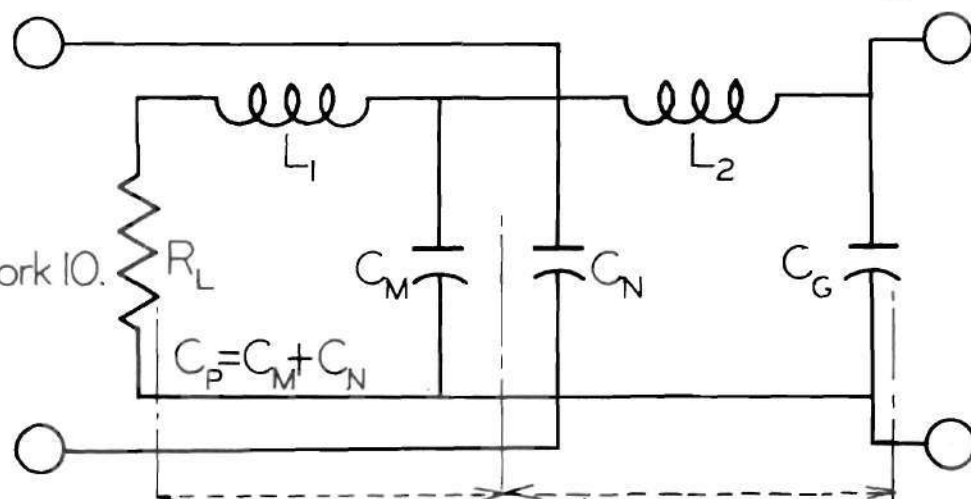
Many of the remarks made in connection with two-terminal filter interstages also apply here. However, in the four-terminal networks it is the filter transfer impedance which must exhibit the characteristics previously described. The transfer impedance may be defined here as the ratio of the signal voltage delivered to the second tube to the plate current of the first tube.

⁴¹Same as footnote 5.

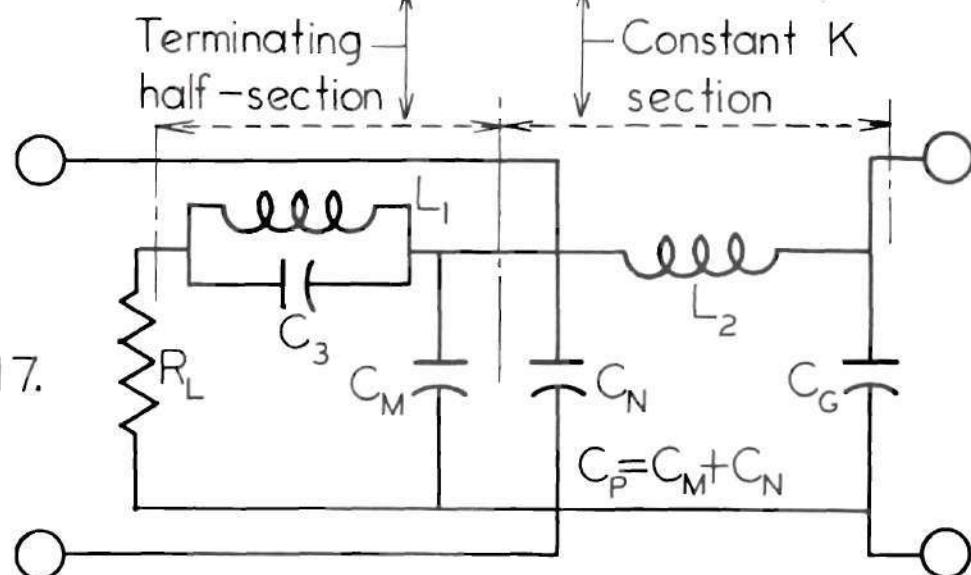
⁴²Bode, op. cit., p. 429.

⁴³Same references as footnote 33. The diagrams of Figures 7(b) and 7(c) were redrawn, with modifications, from Terman's Figures 58(e) and 58(f).

(a) Network 10.



(b) Net. 17.



(c) Net. 18.

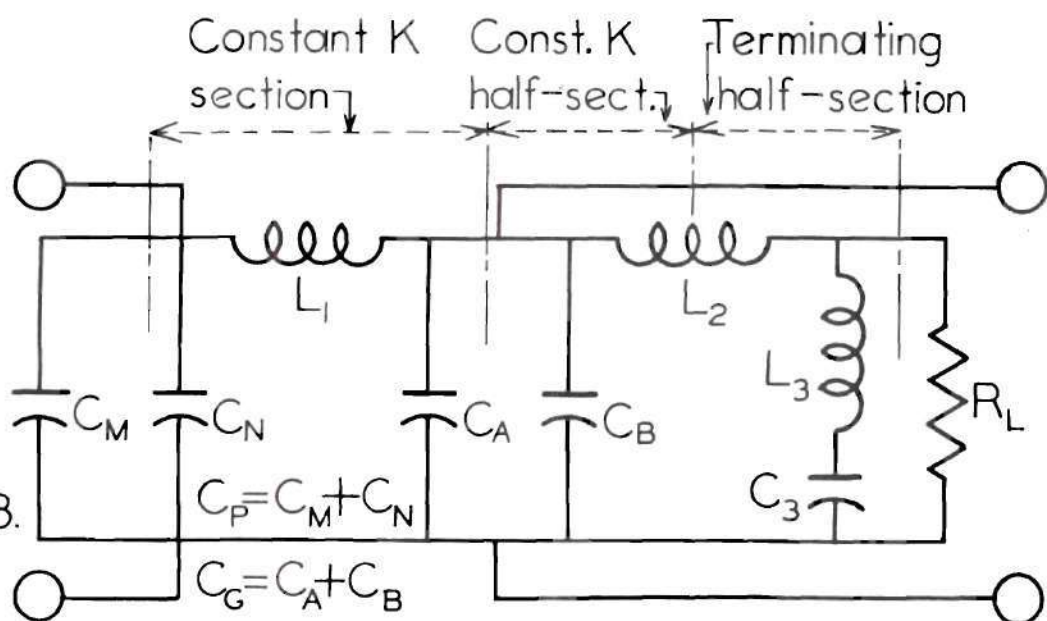


FIG. 7—FOUR-TERMINAL COUPLING NETWORKS SHOWN AS LOW-PASS FILTERS

Modified constant-k sections, properly matched to R_L , are used as in the two-terminal cases. Three of the networks previously described are redrawn as filters in Figure 7 to illustrate the applications of these principles.

Comparison of Performances of Two-Terminal and Four-Terminal Interstages

Wheeler⁴⁴ concludes from filter considerations that the ideal four-terminal interstage gives a band of uniform amplification which is twice as wide as that obtainable with the ideal two-terminal interstage having the same total shunting capacitance. A theorem given in more recent work by Bode⁴⁵ may be interpreted to indicate that a figure of $\pi^2/4$ for relative bandwidth is more plausible.

A four-terminal system gives much more phase shift than a two-terminal system due to the interpolation of a filter section between the tubes in the former.⁴⁶ This is not usually objectionable in itself, the important thing being that the phase shift be proportional to frequency.

The ratio of the capacitances C_p and C_g is an important factor in the design of four-terminal interstages. In most physical circuits the ratio is approximately two, with C_g the larger. Many network designs are based on a C_g to

⁴⁴Wheeler, op. cit.

⁴⁵Bode, op. cit., p. 442

⁴⁶Ibid., p. 428

C_p ratio of precisely two and physical adjustments are sometimes made to produce this condition. Terman⁴⁷ claims that "when $C_g = 2C_p$, the four-terminal system has a possible amplification that is $3/2$ times the theoretical possible gain for a two-terminal network."

Terman also states that the most favorable ratio of capacitances is unity ($C_g = C_p$), resulting in twice as much gain from the four-terminal system as is obtainable with a two-terminal network over the same band. Applying the principle of conservation of bandwidth,⁴⁸ this improvement in performance is seen to be equivalent to the doubling of the bandwidth which was previously mentioned.

It should be pointed out that the conclusions which have been stated here are all expressed in steady-state terms.

It is to be anticipated that, in view of their superior bandwidths, four-terminal interstages will exhibit faster rise-time characteristics than the two-terminal varieties when passing sharp pulses. However, the sharp cut-off characteristics of four-terminal interstages may lead to undesirable transient effects.

⁴⁷Terman, op. cit., p. 421.

⁴⁸This principle amounts to the fact that the product of voltage amplification (numerical ratio) and bandwidth in cycles per second is a constant independent of R_L . See Sarbacher and Edson, op. cit., p. 487.

SQUARE WAVE TESTING

Utility of the Method

The value of square wave test signals in theoretical considerations has already been discussed and it was also pointed out that such signals are very useful in laboratory testing.

Square wave testing is of practical importance from two points of view: (a) It provides a direct measure of the ability of an amplifier to reproduce a sharp pulse; (b) the results may be interpreted in terms of steady-state characteristics.

If steady-state characteristics are of primary interest, as is often the case, it is possible to avoid the tedious process of plotting response curves if one is skilled in the art of square wave testing. Also, since video interstage designs are usually based on steady-state criteria, interpretation of square wave test results in these terms furnishes a correlation between actual performance and the design predictions.

Interpretation of the Results

The information presented here, including Figure 8,*

*The diagrams were redrawn, not reproduced from the article. The captions are essentially direct quotes.

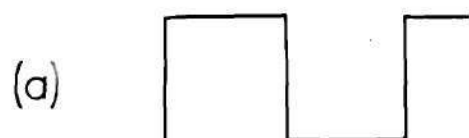
has been taken from an article by Swift.⁴⁹ It is of purely a qualitative nature and is presented as a matter of possible interest to the reader. No attempt will be made here to justify any of the statements which are made.

It has been pointed out that the shape of the output transient at the instant of the application of the input voltage, and for a short time thereafter, is governed by the high-frequency characteristics of the coupling network. The low-frequency behavior determines the value of the output at a much later time.

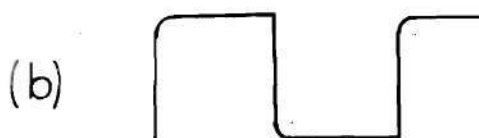
Reference to Figure 8 will clarify the following remarks. In general, a network which passes the high frequencies with very little attenuation will produce an output wave with a sharp leading edge (rapid rise).

The shape of the top of the wave indicates the tendency of the network to oscillate, or the extent of the circuit damping. A circuit which is not heavily damped produces some degree of "overshoot." If the damping is very light, a train of oscillations occurs after the initial rise. The natural frequency of the circuit may be determined approximately by multiplying the frequency of the input signal by the number of maxima which occur in one cycle.

⁴⁹Gilbert Swift, "Amplifier Testing by Means of Square Waves," Communications, 19:22ff., February, 1939.



CONSTANT GAIN AT ALL FREQ.
HIGHLY DAMPED, CONSTANT
DELAY.



SOME ATTENUATION OF HIGH
FREQ. HIGHLY DAMPED,
EXCESS DELAY AT HIGH FREQ.



HIGHER ATTENUATION AT
HIGH FREQUENCIES. HIGHLY
DAMPED, CONSTANT DELAY.



HIGHER ATTENUATION AT
HIGH FREQUENCIES. HIGHLY
DAMPED, EXCESS DELAY AT
HIGH FREQUENCIES.



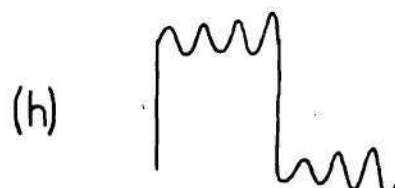
EXCESS GAIN AT HIGH FREQ.
SLIGHT UNDER-DAMPING



SHARP CUT-OFF AT HIGH
FREQ. LESS DAMPING, CONSTANT
DELAY.



EXCESS GAIN AT HIGH FREQ.
PRONOUNCED UNDER-DAMPING.



EXCESS GAIN AT HIGH FREQ.
NEGATIVE DAMPING OR
POSITIVE FEEDBACK.

FIG. 8.—INTERPRETATION OF SQUARE WAVE
RESPONSE IN TERMS OF FREQUENCY AND
PHASE CHARACTERISTICS.

Figure 8(b) shows the effect of excess delay at the higher frequencies. Two diagonally opposite corners are rounded. If the delay is the same for all transmitted frequencies, but the attenuation is high at the upper end of the band, all four corners of the wave have the same shape as illustrated in Figure 8(c).

A circuit whose characteristics include sharp discontinuities will produce trains of oscillations in the output wave. The sharpness of the high-frequency cut-off is indicated by the amplitude and duration of the wave train. The frequency at which the rapid change in the transmission characteristics occurs may be judged from the frequency of the oscillation.

EXPERIMENTAL WORK

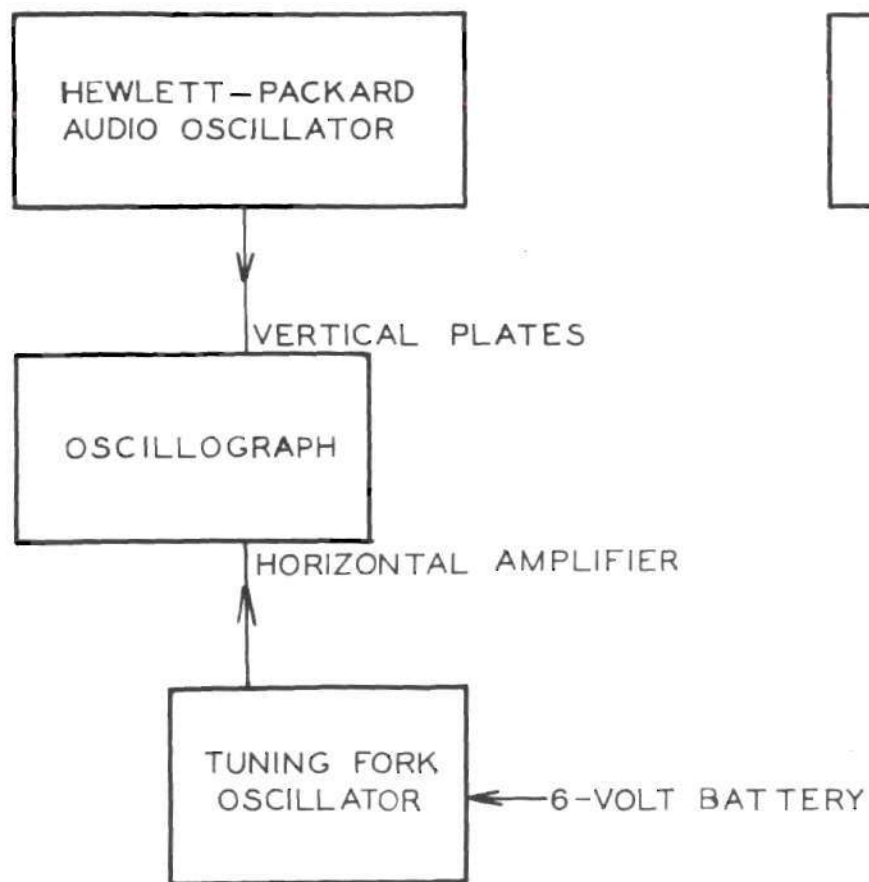
The Apparatus

The arrangement of equipment is shown in Figure 9. A General Radio, 1000 cycle, Type 213 Tuning Fork Audio Oscillator was used as a frequency standard on which to base the tuning of a Hewlett-Packard, Model 200C, Audio Frequency Oscillator. As indicated in Figure 9(a), the Lissajou's figure method of frequency comparison⁵⁰ was used for this purpose.

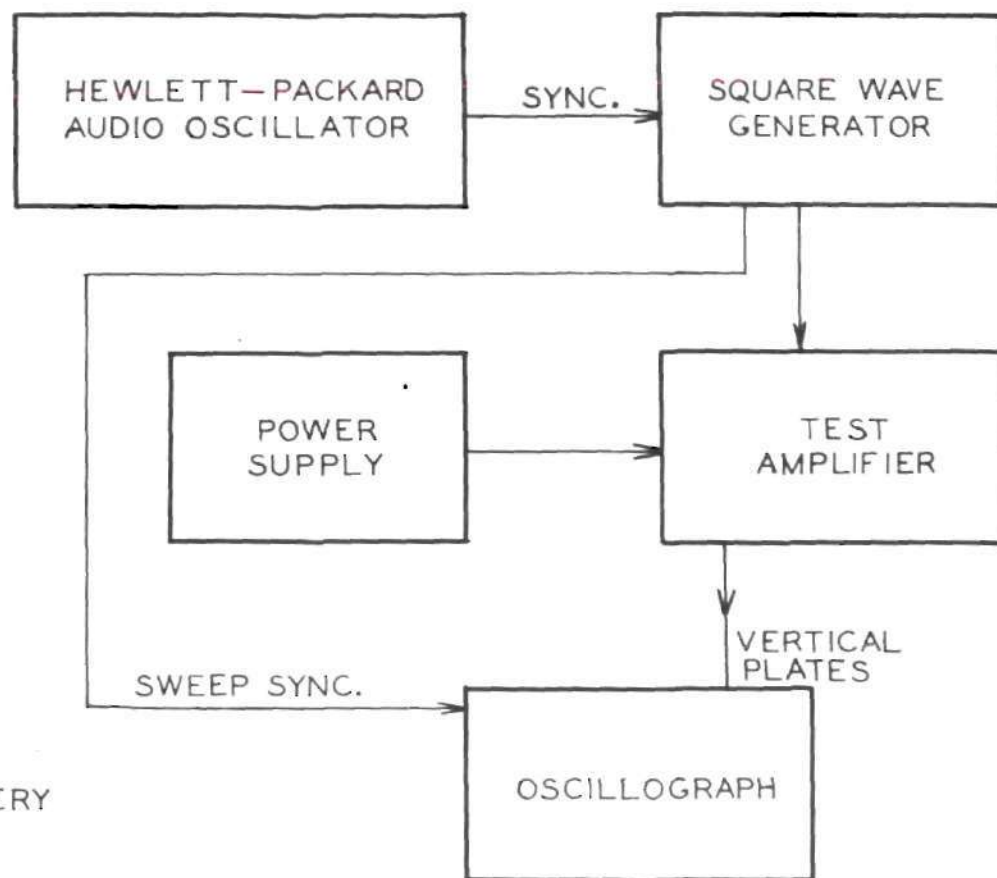
The Hewlett-Packard oscillator supplied a synchronizing signal to a Measurements Corporation, Model 71, Square Wave Generator, as shown in Figure 9(b). The square wave generator supplied signals to an amplifier containing the coupling network under test and the amplifier output was fed directly to the vertical deflecting plates of the cathode-ray tube in a Dumont, Model 208, Oscillograph. The oscillograph sweep was synchronized by the square wave generator output.

The circuit of the test amplifier is shown in Figure 10. It is constructed in such a manner that any type of coupling network can be inserted between the first and second tubes. Four terminals are provided for this purpose

⁵⁰Terman, op. cit., p. 955



(a) Calibration



(b) Tests

FIG.9.- BLOCK DIAGRAMS

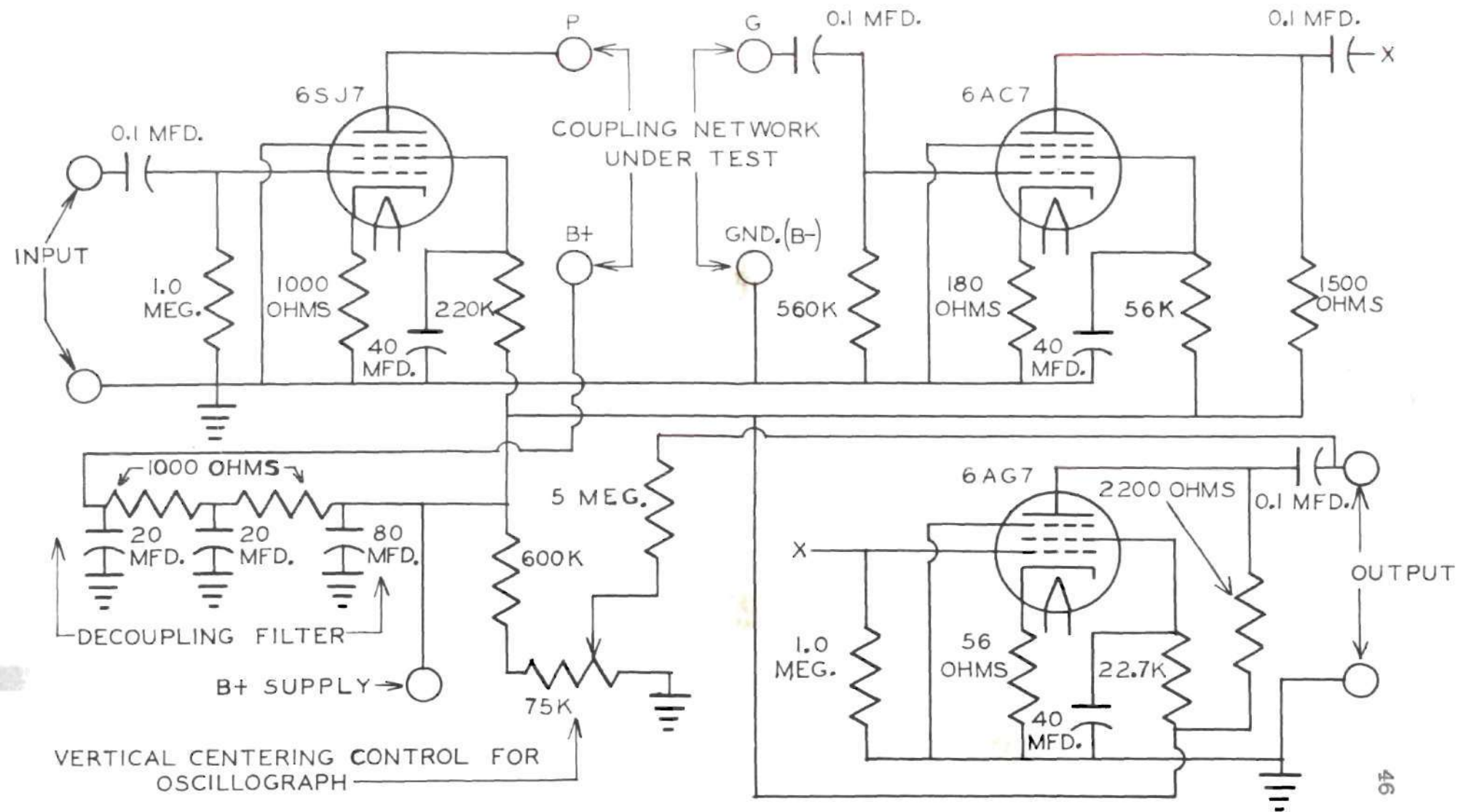


FIG. 10.—TEST AMPLIFIER CIRCUIT

at the points indicated in the circuit diagram. Examples of network connections to these terminals are shown in Figure 11.

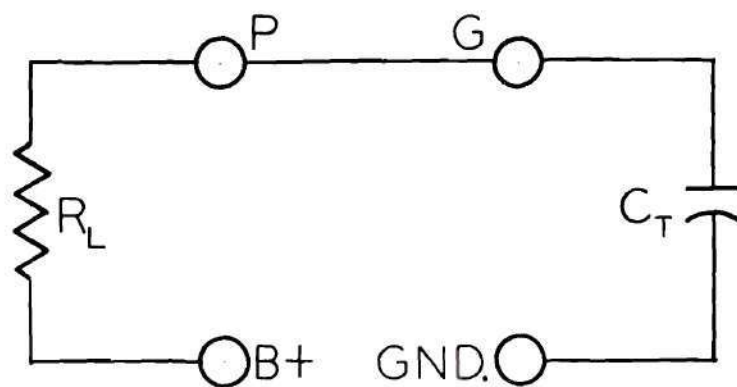
The load resistance, R_L , in the plate circuit of the first tube was kept constant at 1000 ohms during all tests. The values of all plate load resistors were kept as low as possible in order to obtain the best possible high-frequency gain characteristic.⁵¹ The lower limits on the plate load resistors are set by the gain required to produce suitable deflection of the electron beam in the cathode-ray tube. It had been decided to connect the test amplifier output directly to the deflecting plates in order to avoid the distortion which might have been introduced by the vertical amplifier in the oscilloscope. Cathode by-pass condensers were omitted to avoid low-frequency distortion.⁵²

At the time that the equipment was being assembled, the use of a 5000 cycle square wave was anticipated. The vertical amplifier in the oscillograph would not reproduce this wave satisfactorily, but the test amplifier passed it with no observable distortion.

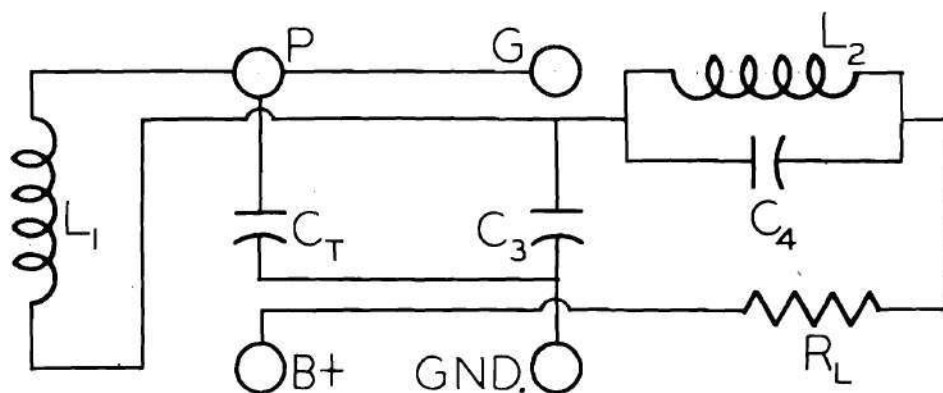
Since the vertical deflecting plates were entirely disconnected from the oscillograph amplifier, the vertical beam centering control was inoperative. Vertical centering

⁵¹Fink, op. cit., p. 214

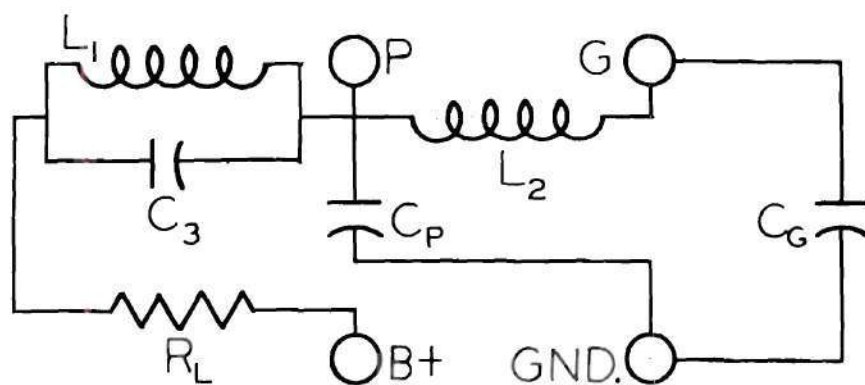
⁵²Ibid., p. 234



(a) Uncompensated interstage



(b) Two-terminal network (7)



(c) Four-terminal network (17)

Note: All circuit components are variable in steps.

FIG.II.—METHODS OF CONNECTING NETWORKS TO AMPLIFIER

is provided in the test amplifier by connecting a variable d-c voltage to one of the output terminals.

The decoupling filter in the plate supply of the first tube was found necessary in order to eliminate oscillations produced by regeneration arising from the common plate load impedance existing between the first and third tubes.⁵³

One of the most difficult problems was the procurement of suitable compensating inductances. After the capabilities of the test amplifier were roughly determined it was possible to select the approximate range of frequencies which would be used. The shunt capacitance to be added to the interstage was determined in a manner to be described later. It was then estimated that four sets of inductances, each consisting of five coils which could be inserted in series in all possible combinations, would be adequate. The values of the five coils in each set were to be 0.5, 1.0, 2.0, 4.0, and 8.0 millihenries, giving a range extending from 0.5 to 15.5 millihenries in 0.5 millihenry steps.

It was essential that the coils have negligible stray capacitance and low resistance. Furthermore the resistance had to be constant over the entire band of frequencies being amplified.

The apparent a-c resistance of a coil is likely to be high and variable with respect to frequency if the coil

⁵³Terman, op. cit., p. 406

has appreciable stray capacitance. This may be explained by the fact that such a coil behaves as a parallel inductance-capacitance circuit having an impedance which is high, particularly at frequencies near resonance.

To minimize the capacitance, pie-wound coils of small physical size were used. Litz wire was used to eliminate the variation of resistance with frequency which would result from skin effect. Powdered iron cores were used to reduce the number of turns required, thereby keeping both physical size and resistance at a minimum. The coils consisted of radio-frequency chokes and intermediate-frequency transformer windings from which turns had been removed to obtain the desired values of inductance. The inductances of the coils were measured, during the turn-removing process, on a General Radio, Type 650-A, 1000 Cycle Impedance Bridge.

The finished coils were assembled in sets and mounted in wooden boxes with a short-circuiting switch shunted across each coil. These switches were of the toggle type and the capacitance across their terminals was of the order of 2 to 3 micro-microfarads as measured with the aid of a Boonton, Type 170-A, Q-Meter. This amount of stray capacitance was insignificant.

The inductances and d-c resistances of the mounted coils were measured on the General Radio 1000 Cycle Impedance Bridge and a Rubicon Portable Wheatstone Bridge respec-

tively. These values are tabulated in Appendix II. The inductances were checked on a Western Electric, Type 4-A, Impedance Bridge at frequencies as high as 20,000 cycles and the values obtained were almost exactly the same as those measured on the other bridge.

Decade condenser boxes in which the smallest increments were 0.001 microfarad were used as the capacitance elements in the networks. The condensers were of the molded mica type with tolerances within 1 per cent.

The load resistance, R_L , was kept constant during the tests. This required the use of a non-inductive resistance which could be decreased in value to compensate for the resistances of coils inserted in the d-c path from B+ to the plate of the tube. The arrangement consisted of two 100 ohm carbon potentiometers and an 820 ohm (nominal) carbon resistor all in series. The combination was calibrated with the Wheatstone bridge. It was possible to keep R_L constant to within about 2 ohms by this means.

An assembly of carbon resistors with a tap switch selecting arrangement was used for the damping resistors R_d and R_s . The nominal values ranged from 100 ohms to 15000 ohms along the logarithmic R.M.A. scale. The resistors were checked on the Wheatstone bridge and only measured values were recorded in the experimental data.

The oscillograms were taken with a Super-Sport Dolly Folding Camera equipped with a Compur Shutter. The lens

speed was $f:2.9$, with a nominal focal length of 7.5 centimeters. The focal length was increased with a 1.2 centimeter extension tube in order to obtain sharp focus at the short distances involved. The camera was mounted in a specially constructed hood. An exposure time of $1/2$ second on Verichrome 120 film at maximum lens opening was adequate.

The Procedure

The square wave frequency to be used in testing a particular network was, in each case, selected so that one half-period would be ten time constants ($R_L C_t$) long. This allowed sufficient time for all important transients to subside before the end of a half-period and still kept the period length well below the upper limit (appearance of low-frequency distortion). The total shunt capacitance, C_t , was the independent variable in the final step of the frequency selection process since R_L was held constant.

The main considerations in the selection of C_t were the anticipated sizes of the required compensating elements. Values calculated in accordance with recommendations taken from the literature were used as first approximations, to be modified by experiment if necessary. Attempts were made to use convenient values, avoiding very small ones as much as possible in the interests of percentage accuracy.

In practice, a tentative frequency was selected and this determined a value for C_t automatically. Theoretical values of the required compensating elements were then

calculated. If the values obtained were not desirable, the assumed frequency was changed and the calculations repeated until the resulting values were satisfactory.

Most of the laboratory manipulations are self-evident from the material which has been presented, but a few points deserve further attention.

All adjustments of frequency, synchronizing voltage, signal amplitude and oscillograph sweep were carried out prior to starting a group of tests. The interstage to be tested was inserted in the amplifier and reduced to simple resistance coupling by appropriate switching. This procedure permitted the observation of the undistorted square wave during the preliminary adjustments.

It was necessary to adjust the equipment to produce a symmetrical* square wave test signal of standard amplitude, the same in all tests. Symmetry was obtained by adjustment of the square wave generator frequency dial and the Hewlett-Packard oscillator output control. The square wave amplitude was adjusted to give a test amplifier output voltage of a value which produced a two inch vertical deflection of the cathode-ray tube.

The oscillograph sweep speed was set so that two or three complete square wave cycles were observed on the

*Positive and negative half-cycles of equal length.

screen. The sweep synchronization was such that the positive half-cycle to be displayed during the tests did not occur at the start of the sweep, but early in the sweep cycle. The horizontal gain control was used to greatly expand this half-cycle to bring it to a standard image size (two inches in width). By these means a fairly small, undistorted portion of the most linear* part of the sweep cycle was employed.

When the adjustments were completed, the tests were run with the network circuit constants first set at the theoretical calculated values. The constants were systematically varied to determine the effect of each separately. Experimental attempts were made to find the most satisfactory combinations of values. Waveforms which seemed to convey significant information were photographed.

Occasional breaks in the oscillogram traces will be noticed. These may be attributed to the presence of four lines which were painted on the face of the cathode-ray tube to serve as guides in the adjustments of image size and position.

*The linearity of the sweep was checked by observing ten cycles of a sine wave and noting whether all half-periods were of equal width on the screen. It was found that at least half of the sweep was linear enough to introduce no observable errors in the results.

Calculated Values of Circuit Constants

The values presented here were calculated in accordance with recommendations of the authors under whose names they are tabulated. In some cases identical (or very nearly identical) sets of values were found in the works of several authors. Only one author is cited here for each set of values, but the others are mentioned in the section headed "Interstage Designs From the Literature," which may be consulted for the actual recommendations and for footnote references to the sources. Sample calculations are presented in Appendix I.

The square wave frequency and total shunt capacitance selected are indicated in each case. $R_L = 1000$ ohms in all cases. Inductances are given in millihenries, capacitances in microfarads, resistances in ohms and frequencies in cycles per second.

Network 1, $f = 2500$, $C_t = 0.02$

Zworykin and Morton

K:	0.50	0.414	0.32	0.37
L:	10.00	8.28	6.40	7.40

Network 2, $f = 2500$, $C_t = 0.02$

No recommendations available, but C_K was assumed to be of the same order of magnitude as C_t .

Network 3, $f = 2500$, $C_t = 0.02$

No recommendations available, but C_K was estimated

as in Network 2. The network is similar to Network 1, so the same orders of magnitude of inductances were assumed.

Network 4, $f = 2500$, $C_t = 0.02$

(a) Terman

$$L = 7.50, \quad C_3 = 0.0133$$

(b) Herold

$$L = 7.60, \quad C_3 = 0.006$$

(c) Bode

$$L = 7.50, \quad C_3 = 0.01335$$

Network 5, $f = 2500$, $C_t = 0.02$

Bode

$$L = 18.00 \text{ or } 17.00, \quad C_3 = 0.01$$

Network 6, $f = 2500$, $C_t = 0.02$

Bode

$$L_1 = 16.00 \text{ or } 15.00, \quad L_2 = 10.67, \quad C_3 = 0.006$$

Network 7, $f = 2500$, $C_t = 0.02$

Bode

$$(a) \quad L_1 = 20.00, \quad L_2 = 6.00, \quad C_3 = 0.016,$$

$$C_4 = 0.01067$$

$$(b) \quad L_1 = 19.314, \quad L_2 = 6.00, \quad C_3 = 0.01593,$$

$$C_4 = 0.008645$$

Network 8, $f = 3000$, $C_t = 0.0167$

Bode

$$L_1 = 14.43, \quad L_2 = 3.89, \quad L_3 = 2.49, \quad L_4 = 2.74,$$

$$C_3 = 0.00851, \quad C_4 = 0.0121$$

Network 9, $f = 2500$, $C_t = 0.02$

(a) Seeley and Kimball

$$C_p = 0.0067, C_g = 0.0133, L = 13.40$$

(b) Herold

$$C_p = 0.004, C_g = 0.016, L = 11.20$$

Network 10, $f = 2000$, $C_t = 0.025$

Seeley and Kimball

$$C_p = 0.0083, C_g = 0.0167, L_1 = 3.00, L_2 = 13.00$$

There were no recommended designs found in the literature for Networks 11 through 16. Each is similar to a previously described network, however, and it was possible to estimate the approximate initial values of circuit elements to use.

Network 17, $f = 2500$, $C_t = 0.02$

Terman

$$C_p = 0.0089, C_g = 0.0111, L_1 = 3.33, L_2 = 11.10,$$

$$C_3 = 0.0059$$

Network 18, $f = 2500$, $C_t = 0.02$

Terman

$$C_p = 0.01, C_g = 0.01, L_1 = 10.00, L_2 = 8.00,$$

$$L_3 = 5.33, C_3 = 0.003$$

RESULTS OF EXPERIMENTS

In this section the experimental data are presented in the form of oscillograms showing the output transients resulting from the application of square wave input signals to the various coupling networks.

The information given on the pages just preceding the oscillograms identifies the figures in terms of the circuit conditions associated with each. An asterisk indicates that the values used were intended to approximate those recommended by one or more of the authors cited in the discussion. These values have been summarized in the preceding section.

Capacitances are expressed in microfarads, inductances in millihenries, resistances in ohms, and input signal frequencies (f) in cycles per second. Resistances are given to the nearest ohm. $R_L = 1000$ ohms in all cases.

2500 Cycle Reference Waveforms:

Fig. 12 - Resistance-coupled amplifier, no added C_t .

Fig. 13 - Resistance-coupled amplifier, $C_t = 0.02$.

Network 1, $f = 2500$, $C_t = 0.02$:

*Fig. 14 - $L = 6.5$

*Fig. 15 - $L = 7.5$

*Fig. 16 - $L = 8.5$

*Fig. 17 - $L = 10.0$

Fig. 18 - $L = 11.0$

Network 2, $f = 2500$, $C_t = 0.02$:

Fig. 19 - $C_k = 0.02$

Fig. 20 - $C_k = 0.022$

Fig. 21 - $C_k = 0.03$

Network 3, $f = 2500$, $C_t = 0.02$:

Fig. 22 - $C_k = 0.02$, $L = 1.0$

Fig. 23 - $C_k = 0.02$, $L = 2.0$

Fig. 24 - $C_k = 0.01$, $L = 7.5$

Fig. 25 - $C_k = 0.015$, $L = 4.5$

Fig. 26 - $C_k = 0.02$, $L = 4$

Network 4, $f = 2500$, $C_t = 0.02$:

*Fig. 27 - $L = 7.5$, $C_3 = 0.013$

Fig. 28 - $L = 7.5$, $C_3 = 0.02$

Fig. 29 - $L = 8.5$, $C_3 = 0.006$

Fig. 30 - $L = 9.5$, $C_3 = 0.006$

*Fig. 31 - $L = 7.5$, $C_3 = 0.006$

Fig. 32 - $L = 6.5$, $C_3 = 0.006$

Network 5, $f = 2500$, $C_t = 0.02$:

*Fig. 33 - $L = 18.0$, $C_3 = 0.01$

*Fig. 34 - $L = 17.0$, $C_3 = 0.01$

Fig. 35 - $L = 16.0$, $C_3 = 0.01$

Fig. 36 - $L = 18.0$, $C_3 = 0.013$

Fig. 37 - $L = 18.0$, $C_3 = 0.007$

Fig. 38 - $L = 12.5$, $C_3 = 0.005$

Network 6, $f = 2500$, $C_t = 0.02$:

*Fig. 39 - $L_1 = 16.0$, $L_2 = 11.0$, $C_3 = 0.006$

Fig. 40 - L_1 increased: $L_1 = 17.0$, $L_2 = 11.0$,
 $C_3 = 0.006$

Fig. 41 - L_1 decreased: $L_1 = 14.0$, $L_2 = 11.0$,
 $C_3 = 0.006$

Fig. 42 - L_2 increased: $L_1 = 16.0$, $L_2 = 14.0$,
 $C_3 = 0.006$

Fig. 43 - L_2 decreased: $L_1 = 16.0$, $L_2 = 8.0$,
 $C_3 = 0.006$

Fig. 44 - C_3 increased: $L_1 = 16.0$, $L_2 = 11.0$,
 $C_3 = 0.007$

Fig. 45 - C_3 decreased: $L_1 = 16.0$, $L_2 = 11.0$,
 $C_3 = 0.005$

Fig. 46 - $L_1 = 13.0$, $L_2 = 8.0$, $C_3 = 0.004$

Network 7, $f = 2500$, $C_t = 0.02$:

*Fig. 47 - $L_1 = 20.0$, $L_2 = 6.0$, $C_3 = 0.016$,
 $C_4 = 0.01$

Fig. 48 - L_1 increased: $L_1 = 22.0$, $L_2 = 6.0$,
 $C_3 = 0.016$, $C_4 = 0.01$

Fig. 49 - L_1 decreased: $L_1 = 18.0$, $L_2 = 6.0$,
 $C_3 = 0.016$, $C_4 = 0.01$

Fig. 50 - L_2 increased: $L_1 = 20.0$, $L_2 = 7.0$,
 $C_3 = 0.016$, $C_4 = 0.01$

Fig. 51 - L_2 decreased: $L_1 = 20.0$, $L_2 = 5.0$,
 $C_3 = 0.016$, $C_4 = 0.01$

Fig. 52 - C_3 increased: $L_1 = 20.0$, $L_2 = 6.0$,
 $C_3 = 0.018$, $C_4 = 0.01$

Fig. 53 - C_3 decreased: $L_1 = 20.0$, $L_2 = 6.0$,
 $C_3 = 0.014$, $C_4 = 0.01$

Fig. 54 - C_4 increased: $L_1 = 20.0$, $L_2 = 6.0$,
 $C_3 = 0.016$, $C_4 = 0.014$

Fig. 55 - C_4 decreased: $L_1 = 20.0$, $L_2 = 6.0$,
 $C_3 = 0.016$, $C_4 = 0.006$

Fig. 56 - $L_1 = 17.0$, $L_2 = 8.0$, $C_3 = 0.016$, $C_4 = 0$

3000 Cycle Reference Waveforms:

Fig. 57 - Resistance-coupled amplifier, no added C_t .

Fig. 58 - Resistance-coupled amplifier, $C_t = 0.017$

Network 8, $f = 3000$, $C_t = 0.017$:

*Fig. 59 - $L_1 = 14.5$, $L_2 = 4.0$, $L_3 = 2.5$, $L_4 = 3.0$,
 $C_3 = 0.008$, $C_4 = 0.012$

Fig. 60 - L_1 increased: $L_1 = 15.5$, $L_2 = 4.0$,
 $L_3 = 2.5$, $L_4 = 3.0$, $C_3 = 0.008$, $C_4 = 0.012$

Fig. 61 - L_1 decreased: $L_1 = 13.5$, $L_2 = 4.0$,
 $L_3 = 2.5$, $L_4 = 3.0$, $C_3 = 0.008$, $C_4 = 0.012$

Fig. 62 - L_2 increased: $L_1 = 14.5$, $L_2 = 6.0$,
 $L_3 = 2.5$, $L_4 = 3.0$, $C_3 = 0.008$, $C_4 = 0.012$

Fig. 63 - L_2 decreased: $L_1 = 14.5$, $L_2 = 2.0$,
 $L_3 = 2.5$, $L_4 = 3.0$, $C_3 = 0.008$, $C_4 = 0.012$

Fig. 64 - L_3 increased: $L_1 = 14.5$, $L_2 = 4.0$,
 $L_3 = 3.5$, $L_4 = 3.0$, $C_3 = 0.008$, $C_4 = 0.012$

Fig. 65 - L_3 decreased: $L_1 = 14.5$, $L_2 = 4.0$,
 $L_3 = 1.5$, $L_4 = 3.0$, $C_3 = 0.008$, $C_4 = 0.012$

- Fig. 66 - L_4 increased: $L_1 = 14.5$, $L_2 = 4.0$,
 $L_3 = 2.5$, $L_4 = 5.0$, $C_3 = 0.008$, $C_4 = 0.012$
- Fig. 67 - L_4 decreased: $L_1 = 14.5$, $L_2 = 4.0$,
 $L_3 = 2.5$, $L_4 = 1.0$, $C_3 = 0.008$, $C_4 = 0.012$
- Fig. 68 - C_3 increased: $L_1 = 14.5$, $L_2 = 4.0$,
 $L_3 = 2.5$, $L_4 = 3.0$, $C_3 = 0.009$, $C_4 = 0.012$
- Fig. 69 - C_3 decreased: $L_1 = 14.5$, $L_2 = 4.0$,
 $L_3 = 2.5$, $L_4 = 3.0$, $C_3 = 0.007$, $C_4 = 0.012$
- Fig. 70 - C_4 increased: $L_1 = 14.5$, $L_2 = 4.0$,
 $L_3 = 2.5$, $L_4 = 3.0$, $C_3 = 0.008$, $C_4 = 0.016$
- Fig. 71 - C_4 decreased: $L_1 = 14.5$, $L_2 = 4.0$,
 $L_3 = 2.5$, $L_4 = 3.0$, $C_3 = 0.008$, $C_4 = 0.008$
- Fig. 72 - $L_1 = 12.0$, $L_2 = 3.5$, $L_3 = 2.5$, $L_4 = 0$,
 $C_3 = 0.007$, $C_4 = 0.01$

Network 9, $f = 2500$, $C_t = 0.02$:

- *Fig. 73 - $C_p = 0.007$, $C_g = 0.013$, $L = 13.5$ ($R_c = 62$)
- Fig. 74 - $C_p = 0.007$, $C_g = 0.013$, $L = 11.0$ ($R_c = 48$)
- Fig. 75 - $C_p = 0.007$, $C_g = 0.013$, $L = 8.0$ ($R_c = 30$)
- Fig. 76 - $C_p = 0.01$, $C_g = 0.01$, $L = 8.0$ ($R_c = 30$)
- Fig. 77 - $C_p = 0.01$, $C_g = 0.01$, $L = 11.0$ ($R_c = 48$)
- *Fig. 78 - $C_p = 0.004$, $C_g = 0.016$, $L = 11.0$ ($R_c = 48$)
- Fig. 79 - $C_p = 0.013$, $C_g = 0.007$, $L = 11.0$ ($R_c = 48$)

2000 Cycle Reference Waveforms:

- Fig. 80 - Resistance-coupled amplifier, no added C_t
- Fig. 81 - Resistance-coupled amplifier, $C_t = 0.025$

Network 10, $f = 2000$, $C_t = 0.025$:

- Fig. 82 - $C_p = 0.008$, $C_g = 0.017$, $L_1 = 3.0$,
 $L_2 = 15.0$ ($R_c = 69$)
- Fig. 83 - $C_p = 0.009$, $C_g = 0.016$, $L_1 = 3.0$,
 $L_2 = 13.0$ ($R_c = 58$)
- Fig. 84 - $C_p = 0.009$, $C_g = 0.016$, $L_1 = 4.0$,
 $L_2 = 12.0$ ($R_c = 51$)
- Fig. 85 - $C_p = 0.012$, $C_g = 0.013$, $L_1 = 3.0$,
 $L_2 = 13.0$ ($R_c = 58$)
- Fig. 86 - $C_p = 0.008$, $C_g = 0.017$, $L_1 = 3.0$,
 $L_2 = 12.0$ ($R_c = 51$)
- Fig. 87 - $C_p = 0.008$, $C_g = 0.017$, $L_1 = 2.0$,
 $L_2 = 13.0$ ($R_c = 58$)
- *Fig. 88 - $C_p = 0.008$, $C_g = 0.017$, $L_1 = 3.0$,
 $L_2 = 13.0$ ($R_c = 58$)
- Fig. 89 - $C_p = 0.008$, $C_g = 0.017$, $L_1 = 5.0$,
 $L_2 = 13.0$ ($R_c = 58$)

Network 11, $f = 2500$, $C_t = 0.02$:

- Fig. 90 - $C_p = 0.007$, $C_g = 0.013$, $L = 13.5$
- Fig. 91 - $C_p = 0.007$, $C_g = 0.013$, $L = 15.5$
- Fig. 92 - $C_p = 0.007$, $C_g = 0.013$, $L = 24.0$
- Fig. 93 - $C_p = 0.01$, $C_g = 0.01$, $L = 24.0$
- Fig. 94 - $C_p = 0.01$, $C_g = 0.01$, $L = 16.0$
- Fig. 95 - $C_p = 0.004$, $C_g = 0.016$, $L = 24.0$

Network 12, $f = 2000$, $C_t = 0.025$:

- Fig. 96 - $C_p = 0.008$, $C_g = 0.017$, $L_1 = 3.0$, $L_2 = 13.0$

- Fig. 97 - $C_p = 0.012$, $C_g = 0.013$, $L_1 = 3.0$, $L_2 = 13.0$
 Fig. 98 - $C_p = 0.012$, $C_g = 0.013$, $L_1 = 7.0$, $L_2 = 13.0$
 Fig. 99 - $C_p = 0.012$, $C_g = 0.013$, $L_1 = 3.0$, $L_2 = 20.0$
 Fig. 100 - $C_p = 0.012$, $C_g = 0.013$, $L_1 = 6.0$, $L_2 = 20.0$
 Fig. 101 - $C_p = 0.008$, $C_g = 0.017$, $L_1 = 6.0$, $L_2 = 25.0$
 Fig. 102 - $C_p = 0.008$, $C_g = 0.017$, $L_1 = 14.0$, $L_2 = 28.0$

Network 13, $f = 2500$, $C_t = 0.02$:

- Fig. 103 - $C_p = 0.007$, $C_g = 0.013$, $L = 13.5$, $R_s = 104$
 $(R_c = 62)$
 Fig. 104 - $C_p = 0.007$, $C_g = 0.013$, $L = 15.5$, $R_s = 235$
 $(R_c = 74)$
 Fig. 105 - $C_p = 0.007$, $C_g = 0.013$, $L = 18.5$, $R_s = 392$
 $(R_c = 95)$
 Fig. 106 - $C_p = 0.01$, $C_g = 0.01$, $L = 13.5$, $R_s = 235$
 $(R_c = 62)$
 Fig. 107 - $C_p = 0.01$, $C_g = 0.01$, $L = 18.5$, $R_s = 488$
 $(R_c = 95)$
 Fig. 108 - $C_p = 0.01$, $C_g = 0.01$, $L = 23.5$, $R_s = 792$
 $(R_c = 98)$

Network 14, $f = 2500$, $C_t = 0.02$:

- Fig. 109 - $C_p = 0.007$, $C_g = 0.013$, $L = 13.5$,
 $R_d = 6920$ ($R_c = 62$)
 Fig. 110 - $C_p = 0.007$, $C_g = 0.013$, $L = 15.5$,
 $R_d = 2980$ ($R_c = 74$)
 Fig. 111 - $C_p = 0.007$, $C_g = 0.013$, $L = 13.5$,
 $R_d = 4920$ ($R_c = 62$)

Fig.112 - $C_p = 0.01$, $C_g = 0.01$, $L = 13.5$,
 $R_d = 4920$ ($R_c = 62$)

Fig.113 - $C_p = 0.01$, $C_g = 0.01$, $L = 15.5$,
 $R_d = 3580$ ($R_c = 74$)

Fig.114 - $C_p = 0.01$, $C_g = 0.01$, $L = 18.5$,
 $R_d = 2095$ ($R_c = 95$)

Network 15, $f = 2000$, $C_t = 0.025$:

Fig.115 - $C_p = 0.008$, $C_g = 0.017$, $L_1 = 3.0$,
 $L_2 = 15.0$, $R_s = 104$ ($R_c = 69$)

Fig.116 - $C_p = 0.008$, $C_g = 0.017$, $L_1 = 3.0$,
 $L_2 = 17.0$, $R_s = 235$ ($R_c = 83$)

Fig.117 - $C_p = 0.012$, $C_g = 0.013$, $L_1 = 3.0$,
 $L_2 = 13.0$, $R_s = 104$ ($R_c = 57$)

Fig.118 - $C_p = 0.012$, $C_g = 0.013$, $L_1 = 3.0$,
 $L_2 = 18.0$, $R_s = 336$ ($R_c = 91$)

Fig.119 - $C_p = 0.012$, $C_g = 0.013$, $L_1 = 5.0$,
 $L_2 = 18.0$, $R_s = 392$ ($R_c = 91$)

Fig.120 - $C_p = 0.012$, $C_g = 0.013$, $L_1 = 6.0$,
 $L_2 = 21.0$, $R_s = 607$ ($R_c = 100$)

Fig.121 - $C_p = 0.012$, $C_g = 0.013$, $L_1 = 8.0$,
 $L_2 = 20.0$, $R_s = 647$ ($R_c = 94$)

Fig.122 - $C_p = 0.012$, $C_g = 0.013$, $L_1 = 6.0$,
 $L_2 = 23.0$, $R_s = 647$ ($R_c = 94$)

Network 16, $f = 2000$, $C_t = 0.025$:

Fig.123 - $C_p = 0.008$, $C_g = 0.017$, $L_1 = 3.0$,
 $L_2 = 15.0$, $R_d = 4920$ ($R_c = 69$)

Fig.124 - $C_p = 0.012$, $C_g = 0.013$, $L_1 = 4.0$,

$L_2 = 13.0$, $R_d = 7400$ ($R_c = 57$)

Fig.125 - $C_p = 0.012$, $C_g = 0.013$, $L_1 = 3.0$,

$L_2 = 15.0$, $R_d = 3297$ ($R_c = 69$)

Fig.126 - $C_p = 0.012$, $C_g = 0.013$, $L_1 = 3.0$,

$L_2 = 19.0$, $R_d = 2095$ ($R_c = 86$)

Network 17, $f = 2500$, $C_t = 0.02$:

*Fig.127 - $C_p = 0.009$, $C_g = 0.011$, $L_1 = 3.0$,

$L_2 = 11.0$, $C_3 = 0.006$

Fig.128 - C_3 increased: $C_p = 0.009$, $C_g = 0.011$,

$L_1 = 3.0$, $L_2 = 11.0$, $C_3 = 0.007$

Fig.129 - C_3 decreased: $C_p = 0.009$, $C_g = 0.011$,

$L_1 = 3.0$, $L_2 = 11.0$, $C_3 = 0.004$

Fig.130 - L_1 increased: $C_p = 0.009$, $C_g = 0.011$,

$L_1 = 3.5$, $L_2 = 11.0$, $C_3 = 0.006$

Fig.131 - L_1 decreased: $C_p = 0.009$, $C_g = 0.011$,

$L_1 = 2.5$, $L_2 = 11.0$, $C_3 = 0.006$

Fig.132 - L_2 increased: $C_p = 0.009$, $C_g = 0.011$,

$L_1 = 3.0$, $L_2 = 13.0$, $C_3 = 0.006$

Fig.133 - L_2 decreased: $C_p = 0.009$, $C_g = 0.011$,

$L_1 = 3.0$, $L_2 = 9.0$, $C_3 = 0.006$

Fig.134 - $C_p = 0.01$, $C_g = 0.01$, $L_1 = 3.0$, $L_2 = 11.0$,

$C_3 = 0.006$

Fig.135 - $C_p = 0.007$, $C_g = 0.013$, $L_1 = 3.0$,

$L_2 = 11.0$, $C_3 = 0.006$

Fig.136 - $C_p = 0.007$, $C_g = 0.013$, $L_1 = 2.5$,

$L_2 = 11.0$, $C_3 = 0.006$

Fig.137 - $C_p = 0.007$, $C_g = 0.013$, $L_1 = 2.5$,
 $L_2 = 11.0$, $C_3 = 0.004$

Network 18, $f = 2500$, $C_t = 0.02$

*Fig.138 - $C_p = 0.01$, $C_g = 0.01$, $L_1 = 10.0$,
 $L_2 = 8.0$, $L_3 = 5.5$, $C_3 = 0.003$

Fig.139 - L_1 increased: $C_p = 0.01$, $C_g = 0.01$,
 $L_1 = 12.0$, $L_2 = 8.0$, $L_3 = 5.5$, $C_3 = 0.003$

Fig.140 - L_1 decreased: $C_p = 0.01$, $C_g = 0.01$,
 $L_1 = 9.0$, $L_2 = 8.0$, $L_3 = 5.5$, $C_3 = 0.003$

Fig.141 - L_2 increased: $C_p = 0.01$, $C_g = 0.01$,
 $L_1 = 10.0$, $L_2 = 9.0$, $L_3 = 5.5$, $C_3 = 0.003$

Fig.142 - L_2 decreased: $C_p = 0.01$, $C_g = 0.01$,
 $L_1 = 10.0$, $L_2 = 7.0$, $L_3 = 5.5$, $C_3 = 0.003$

Fig.143 - L_3 decreased: $C_p = 0.01$, $C_g = 0.01$,
 $L_1 = 10.0$, $L_2 = 8.0$, $L_3 = 4.0$, $C_3 = 0.003$

Fig.144 - L_3 increased: $C_p = 0.01$, $C_g = 0.01$,
 $L_1 = 10.0$, $L_2 = 8.0$, $L_3 = 6.5$, $C_3 = 0.003$

Fig.145 - C_3 increased: $C_p = 0.01$, $C_g = 0.01$,
 $L_1 = 10.0$, $L_2 = 8.0$, $L_3 = 5.5$, $C_3 = 0.004$

Fig.146 - C_3 decreased: $C_p = 0.01$, $C_g = 0.01$,
 $L_1 = 10.0$, $L_2 = 8.0$, $L_3 = 5.5$, $C_3 = 0.002$

Fig.147 - $C_p = 0.007$, $C_g = 0.013$, $L_1 = 10.0$,
 $L_2 = 8.0$, $L_3 = 5.5$, $C_3 = 0.003$

Fig.148 - $C_p = 0.01$, $C_g = 0.01$, $L_1 = 8.0$, $L_2 = 6.0$,
 $L_3 = 3.0$, $C_3 = 0.003$

Fig.149 - $C_p = 0.007$, $C_g = 0.013$, $L_1 = 8.5$,
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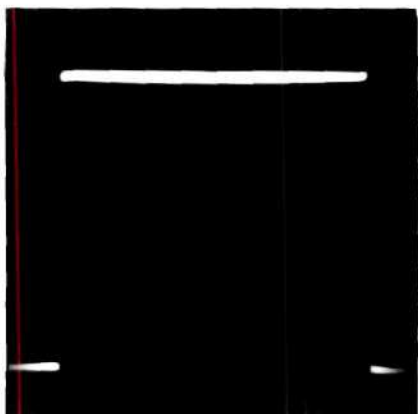


Fig. 12

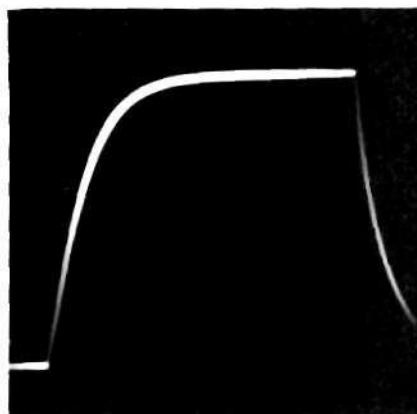


Fig. 13

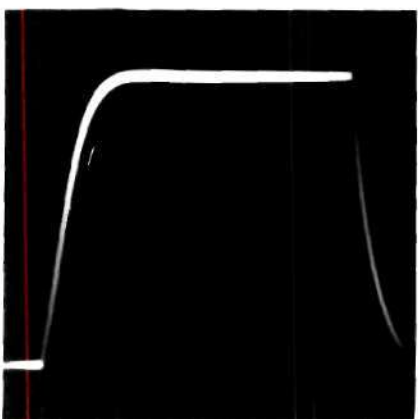


Fig. 14

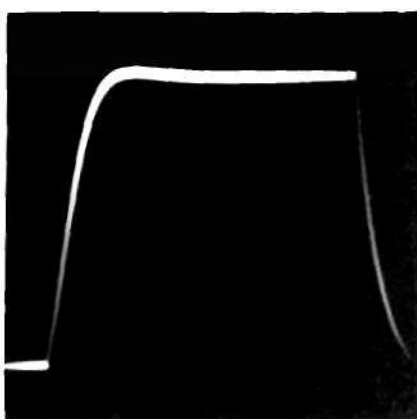


Fig. 15

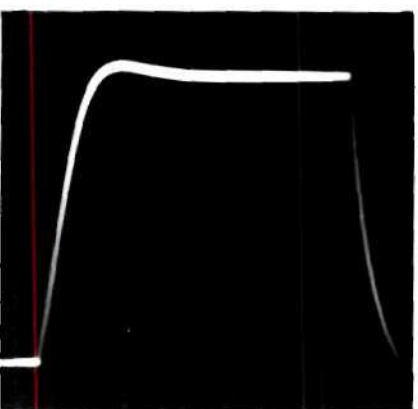


Fig. 16

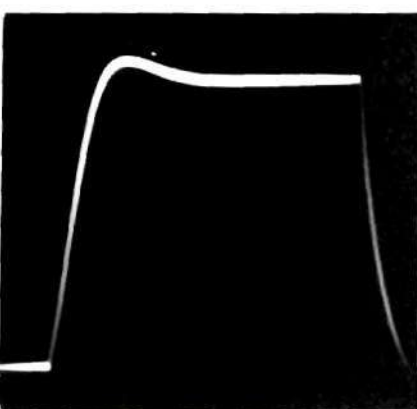


Fig. 17

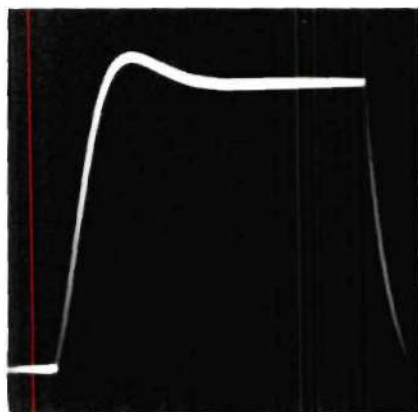


Fig. 18

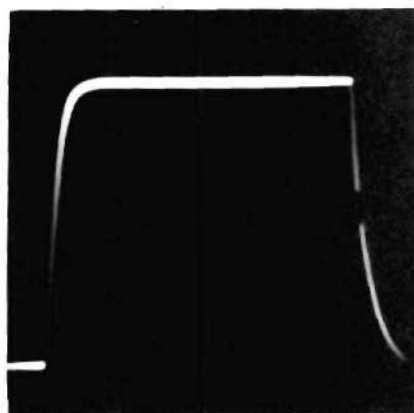


Fig. 19

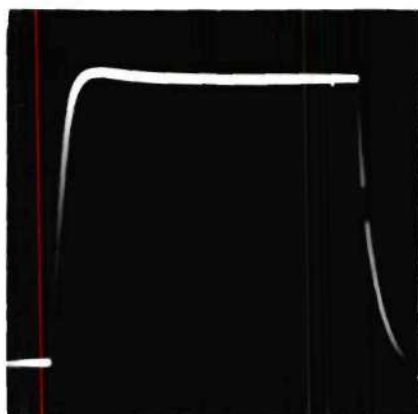


Fig. 20

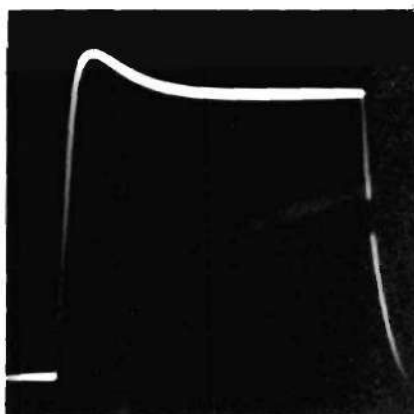


Fig. 21

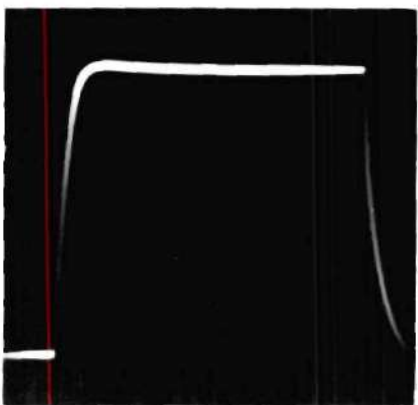


Fig. 22

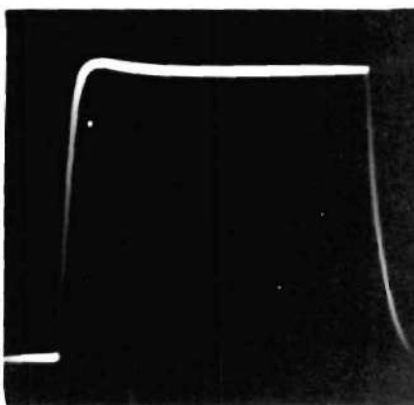


Fig. 23

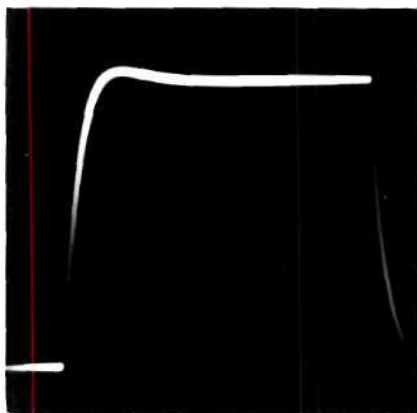


Fig. 24

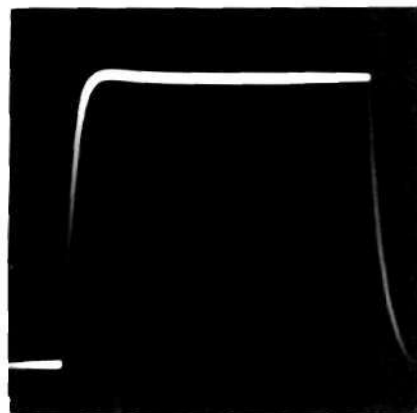


Fig. 25

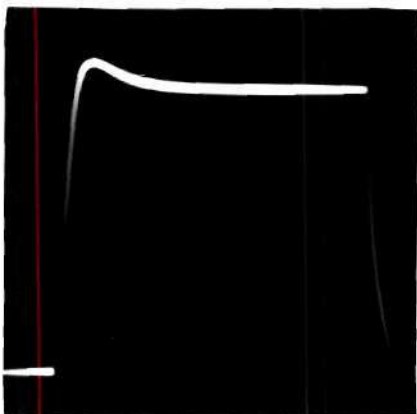


Fig. 26

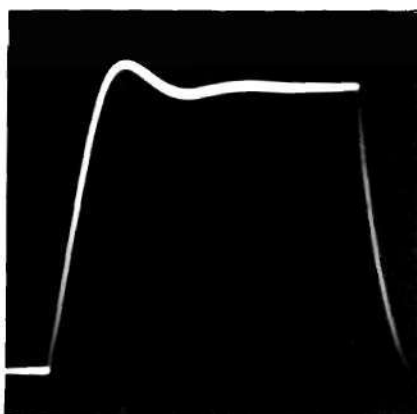


Fig. 27

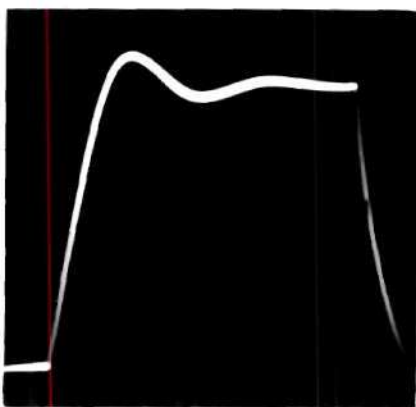


Fig. 28

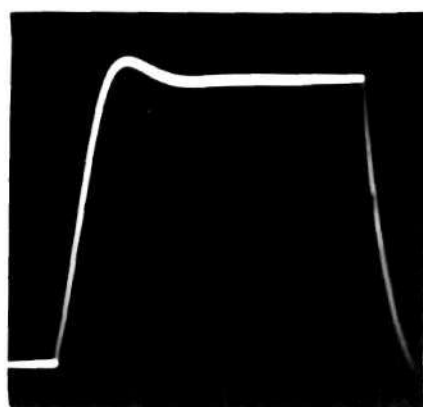


Fig. 29

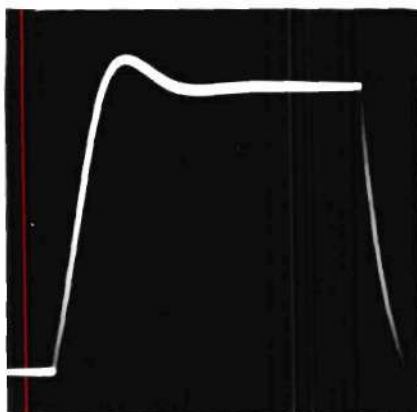


Fig. 30

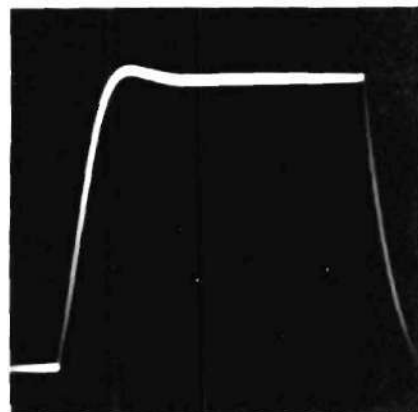


Fig. 31

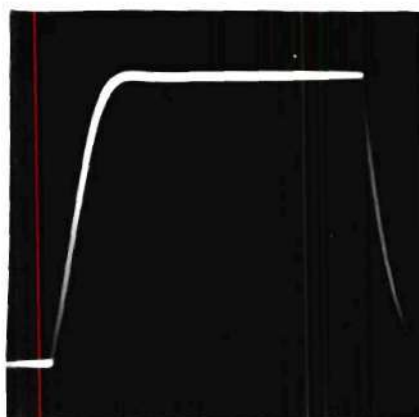


Fig. 32

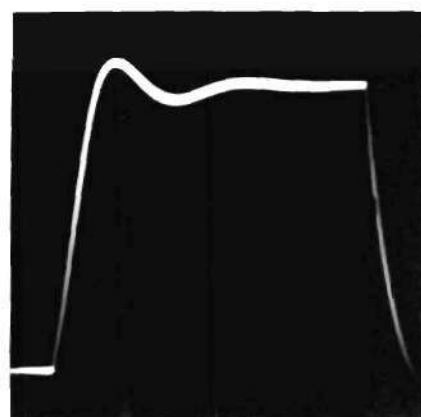


Fig. 33

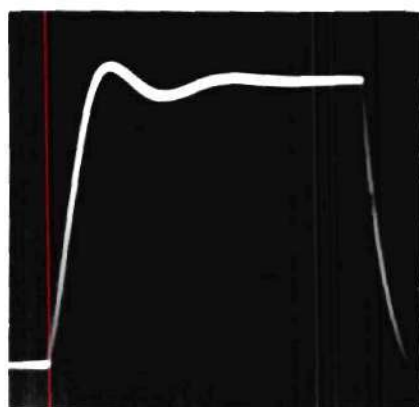


Fig. 34

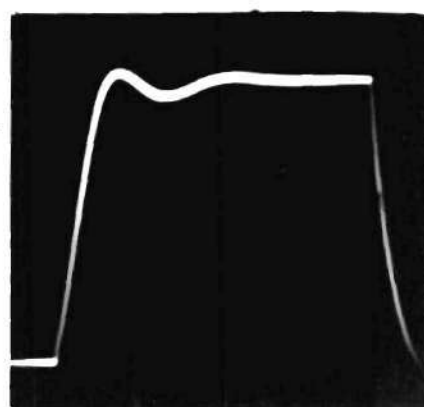


Fig. 35

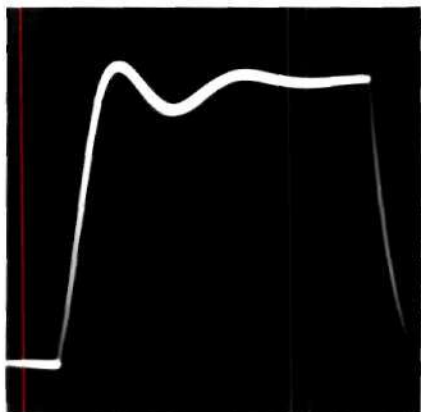


Fig. 36

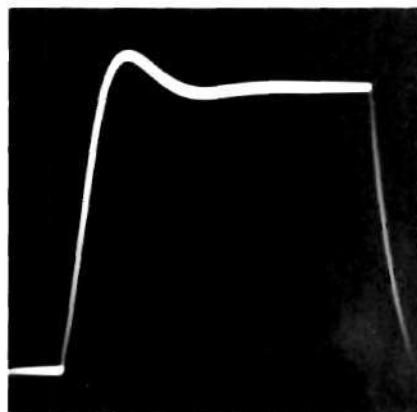


Fig. 37

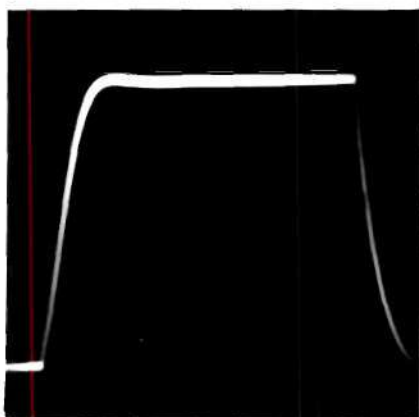


Fig. 38

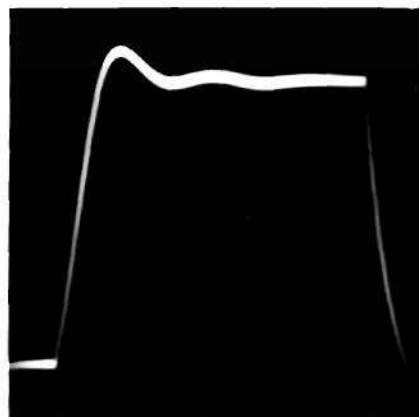


Fig. 39

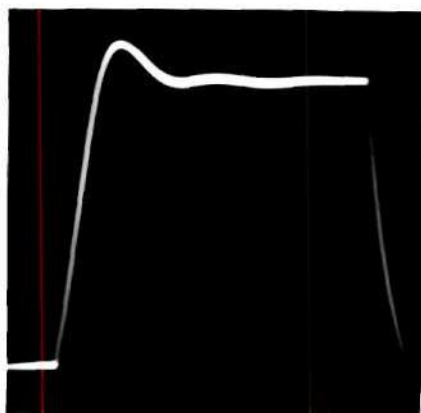


Fig. 40

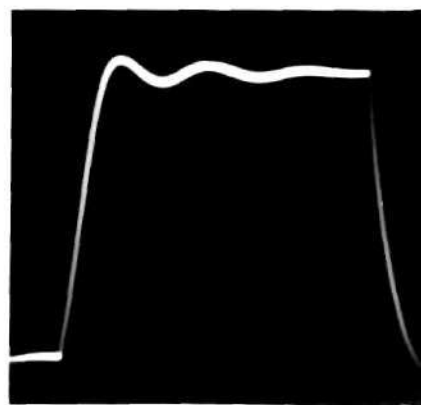


Fig. 41

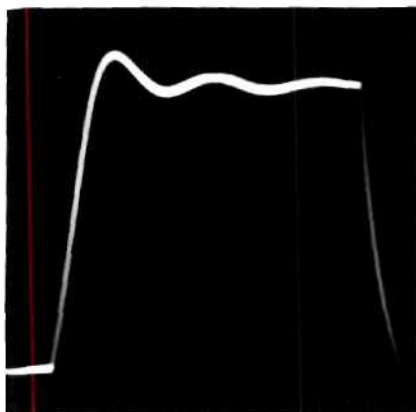


Fig. 42

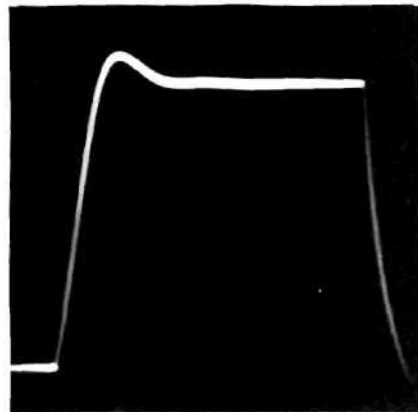


Fig. 43

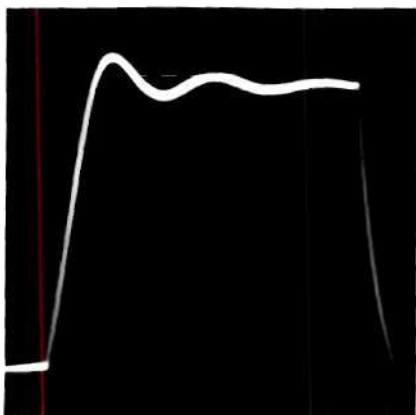


Fig. 44

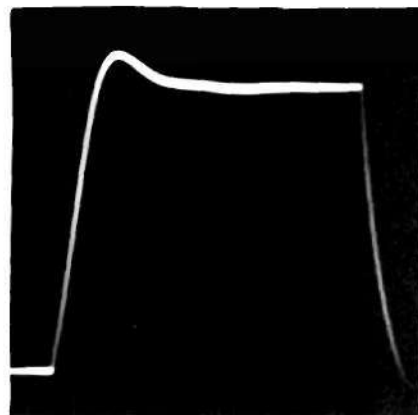


Fig. 45

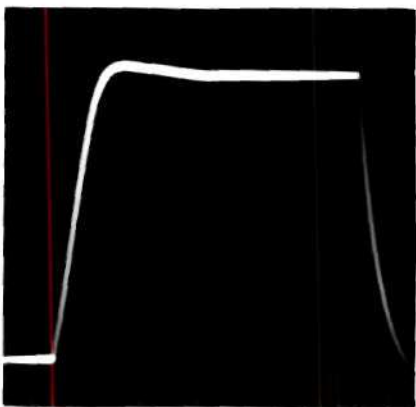


Fig. 46

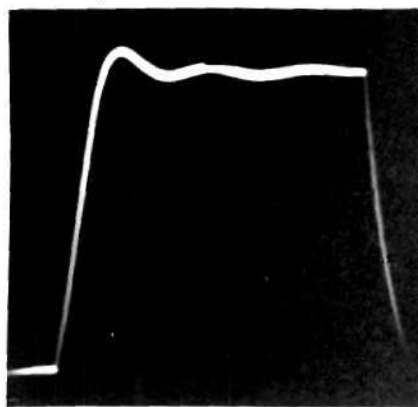


Fig. 47

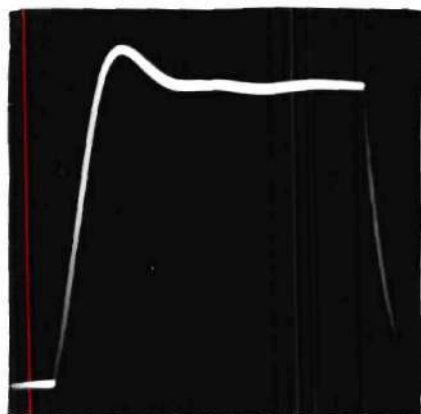


Fig. 48

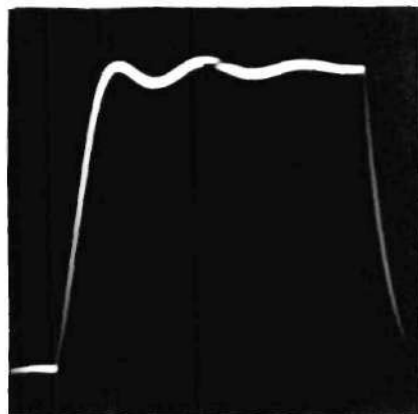


Fig. 49

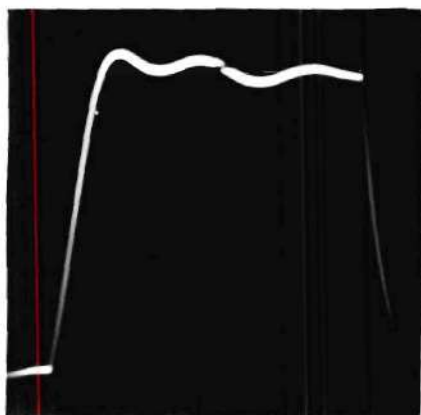


Fig. 50

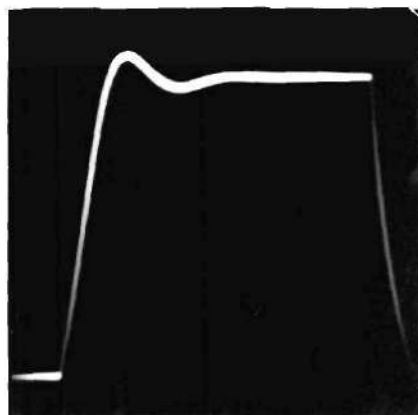


Fig. 51

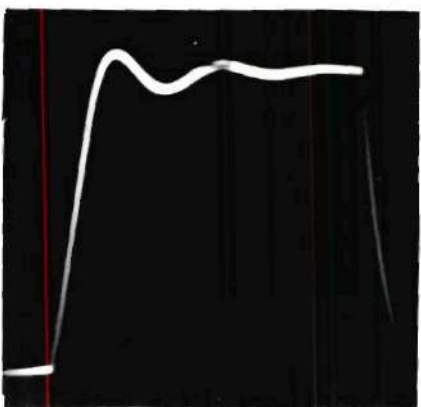


Fig. 52

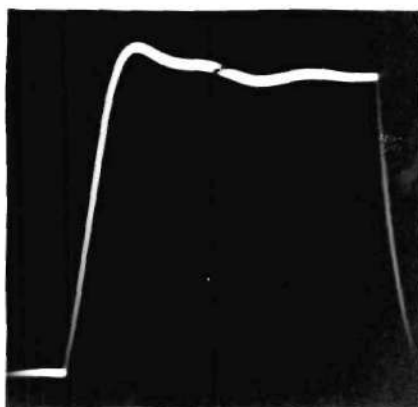


Fig. 53

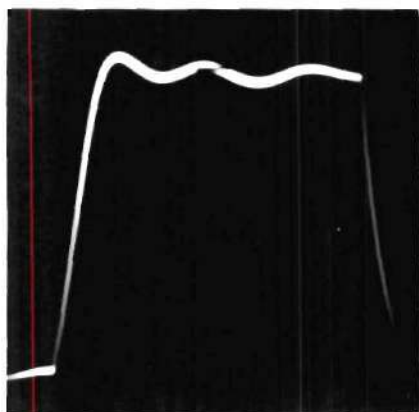


Fig. 54

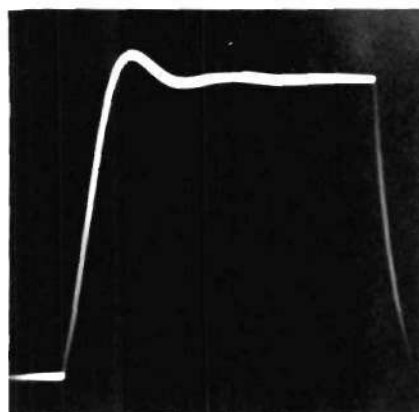


Fig. 55

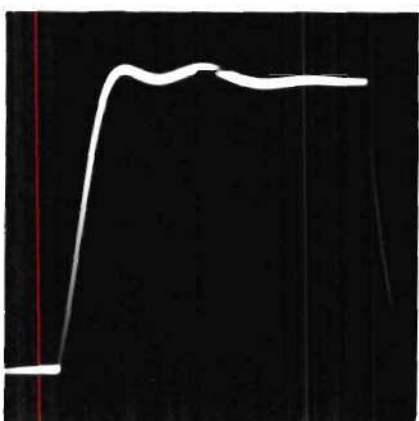


Fig. 56

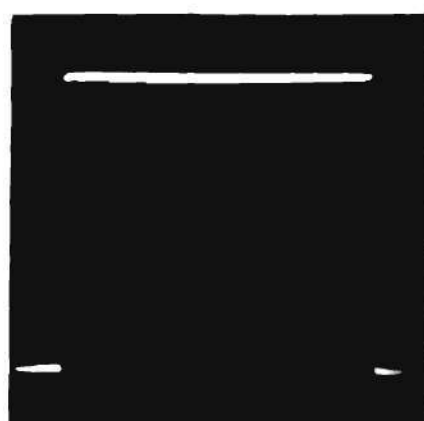


Fig. 57

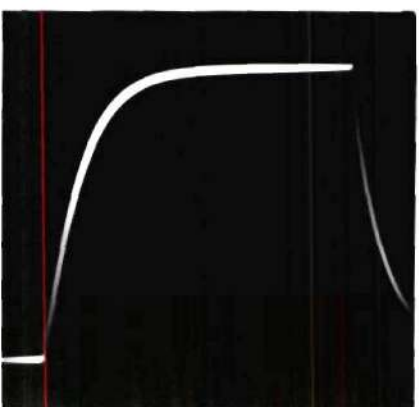


Fig. 58

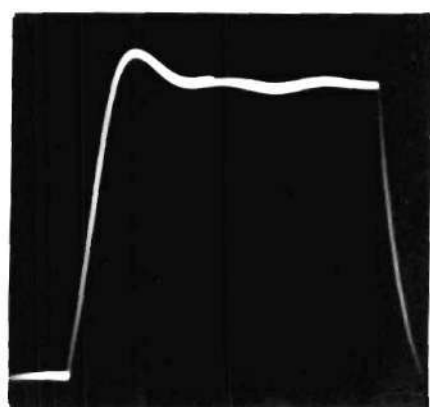


Fig. 59

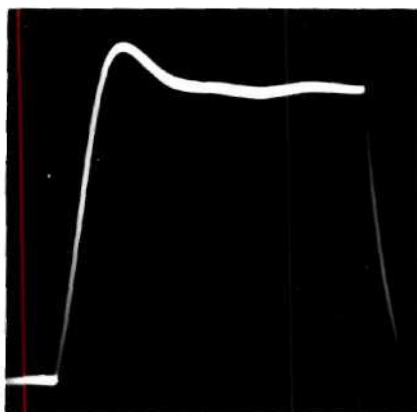


Fig. 60

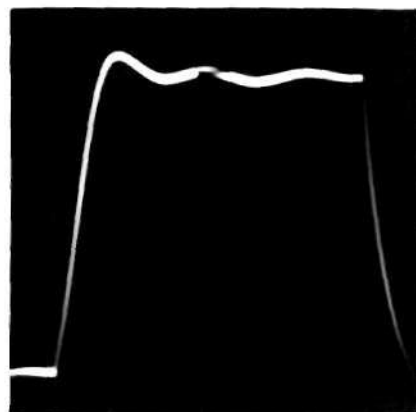


Fig. 61

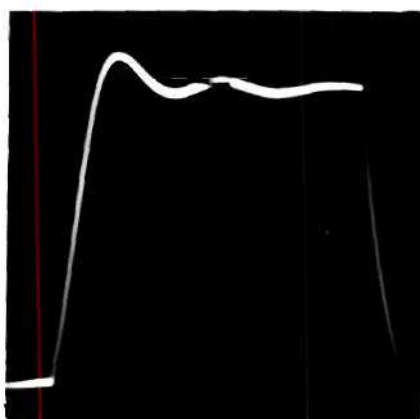


Fig. 62

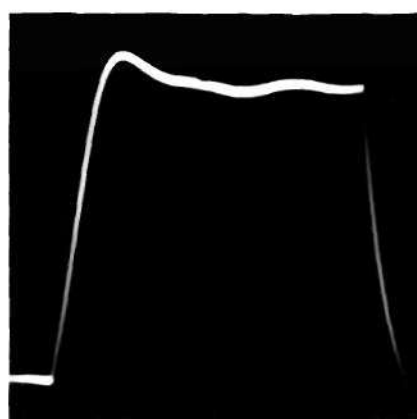


Fig. 63

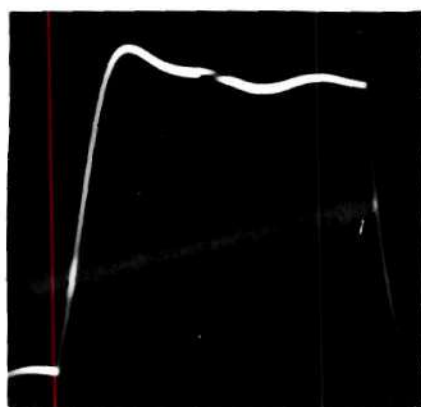


Fig. 64

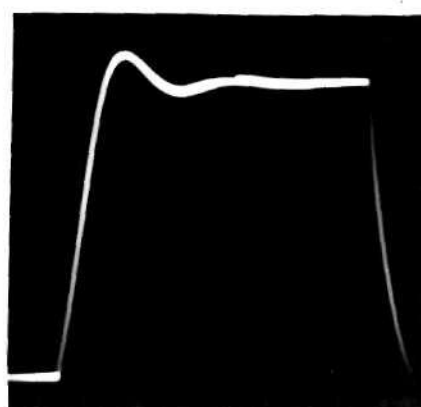


Fig. 65

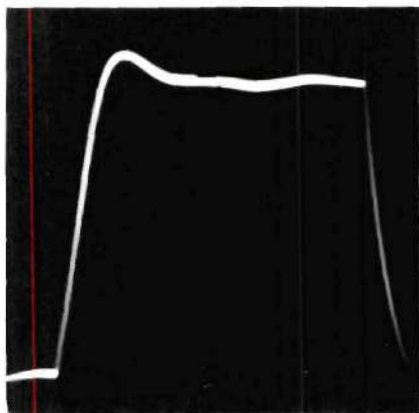


Fig. 66

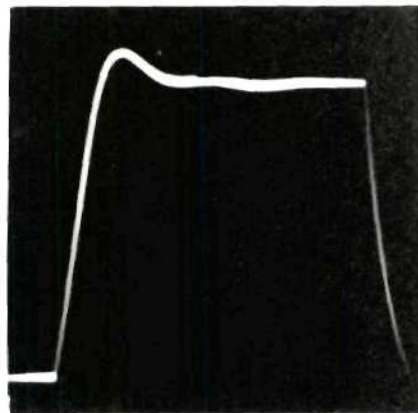


Fig. 67

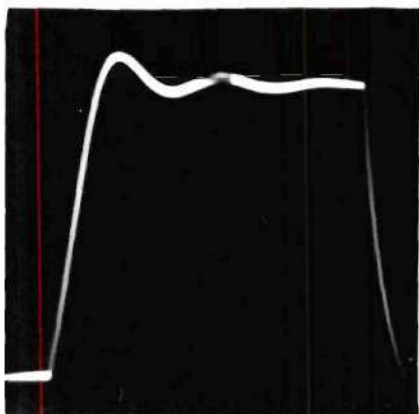


Fig. 68

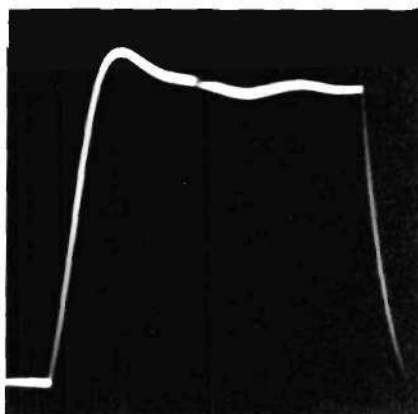


Fig. 69

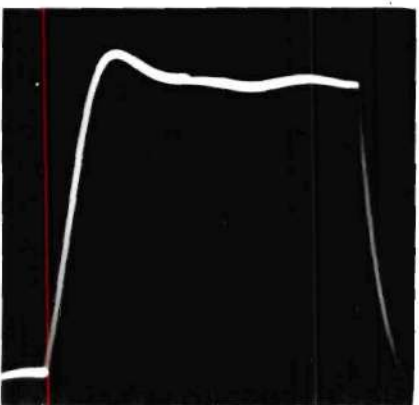


Fig. 70

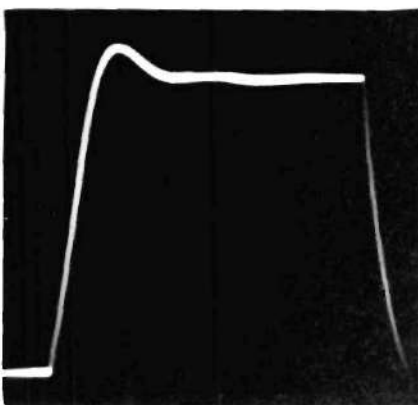


Fig. 71

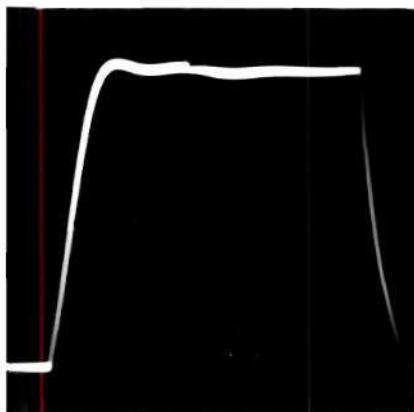


Fig. 72

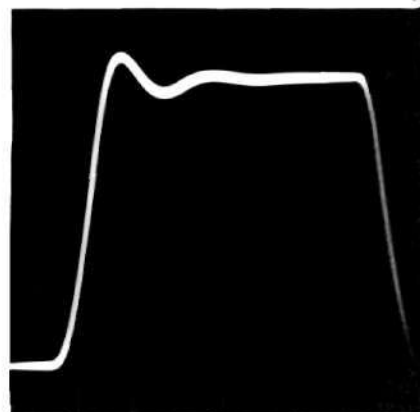


Fig. 73

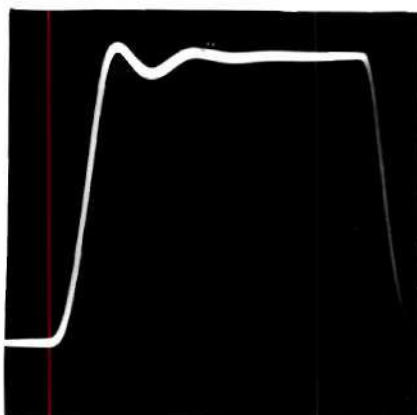


Fig. 74

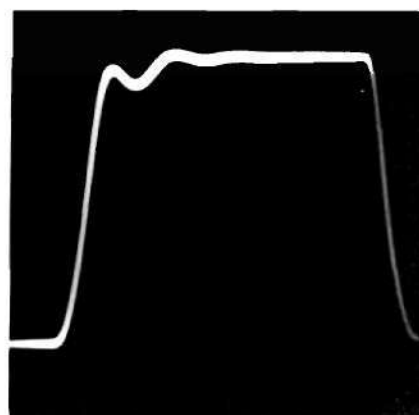


Fig. 75

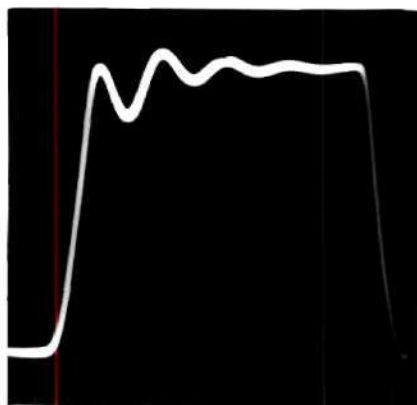


Fig. 76

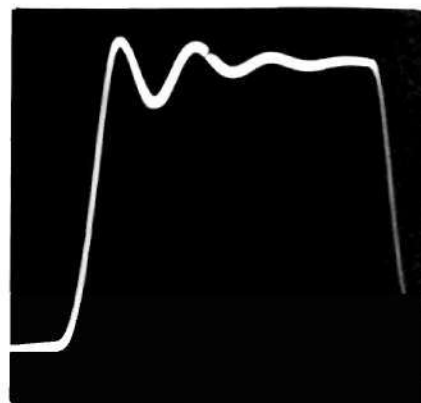


Fig. 77

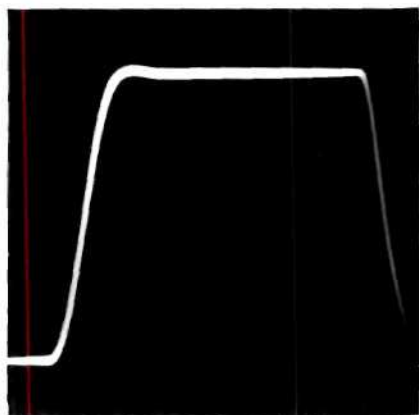


Fig. 78

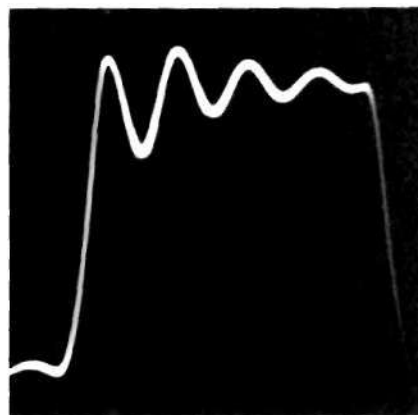


Fig. 79

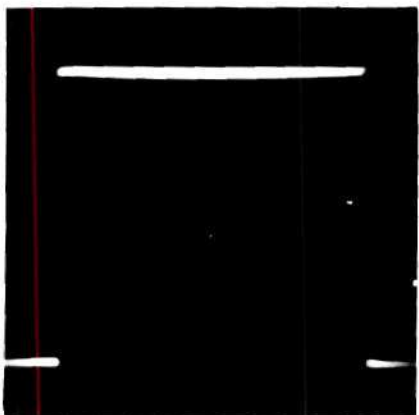


Fig. 80

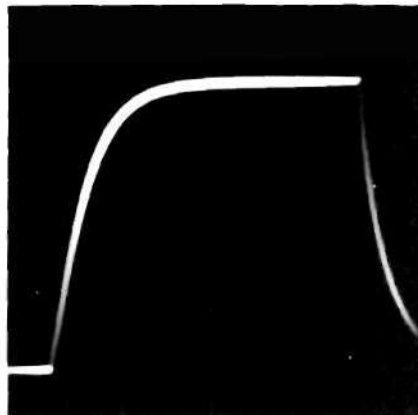


Fig. 81

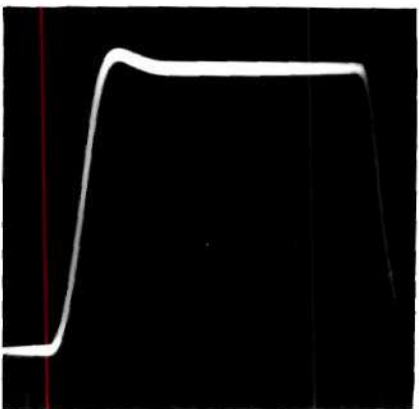


Fig. 82

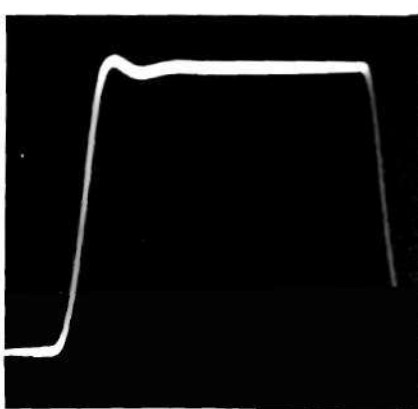


Fig. 83

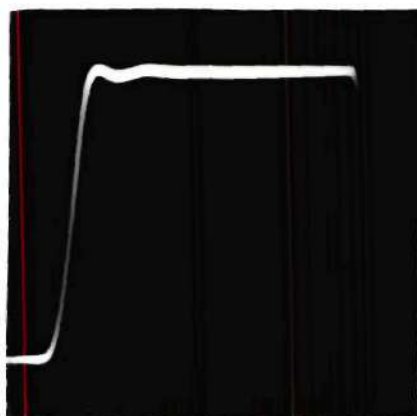


Fig. 84

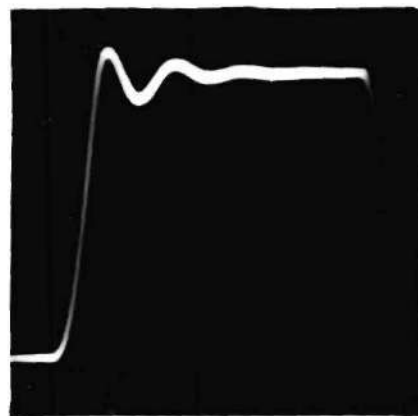


Fig. 85

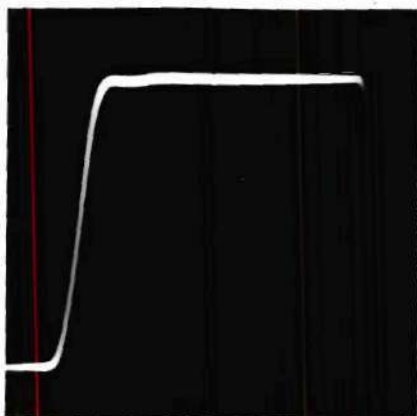


Fig. 86

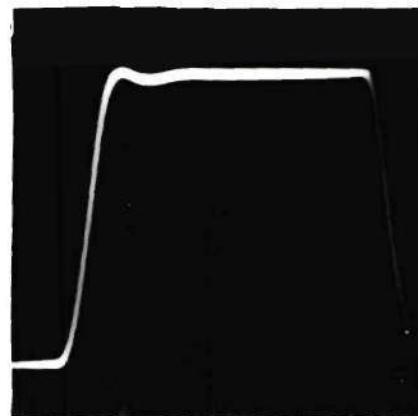


Fig. 87

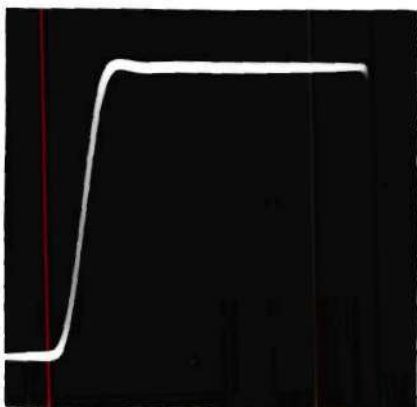


Fig. 88



Fig. 89

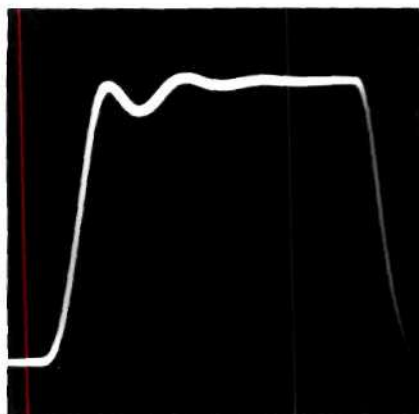


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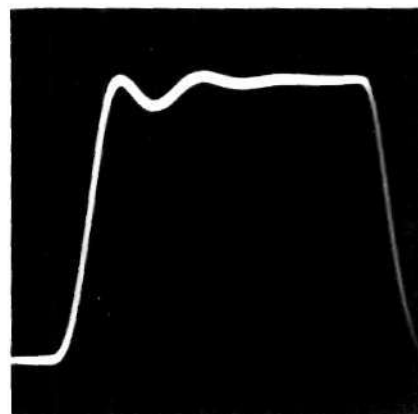


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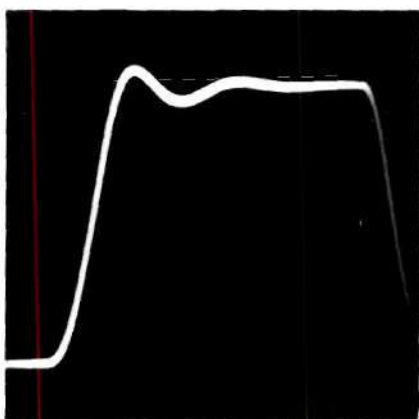


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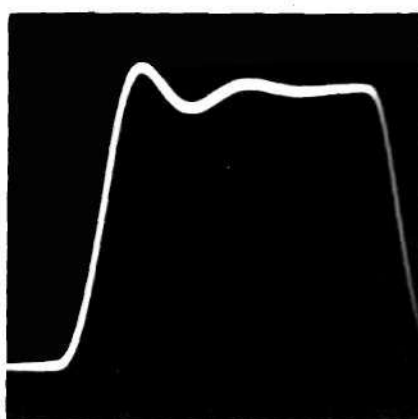


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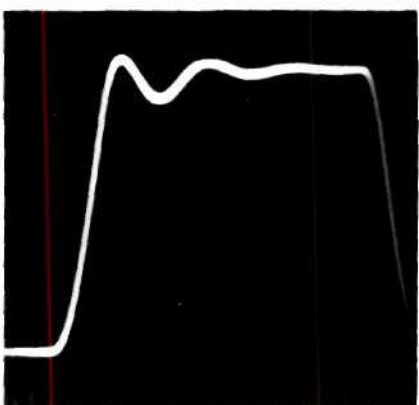


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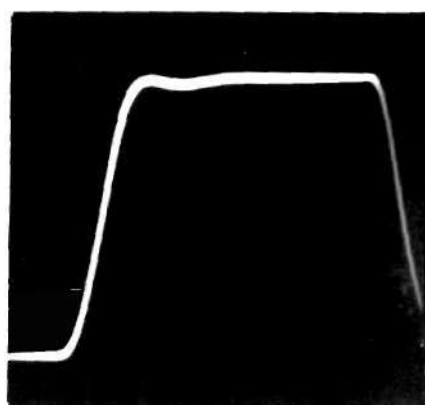


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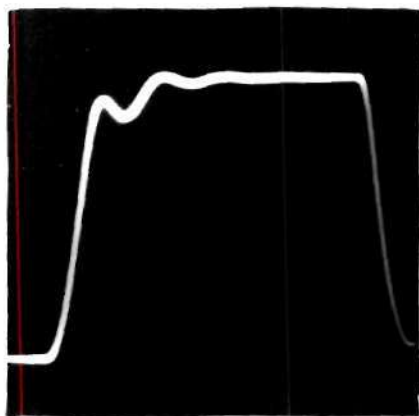


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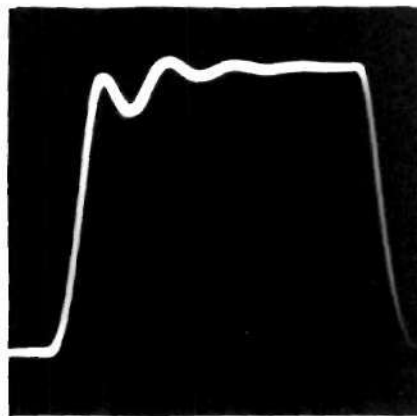


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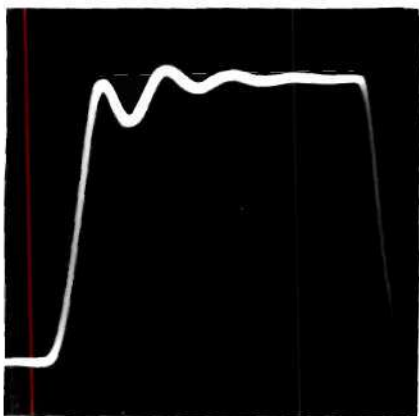


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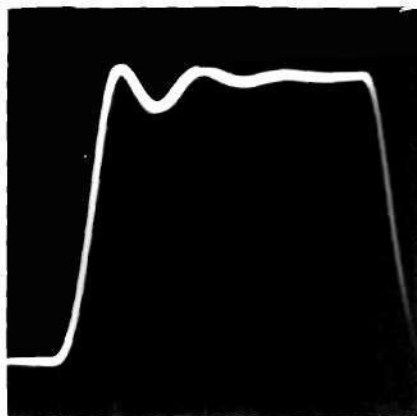


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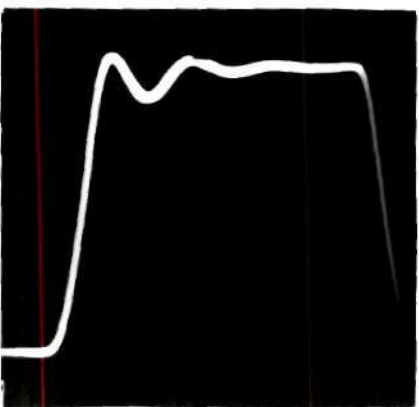


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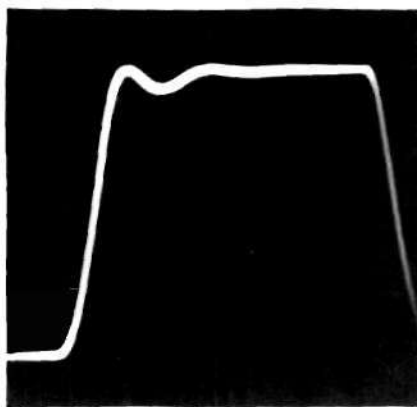


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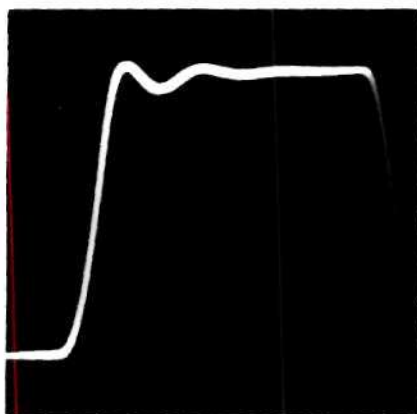


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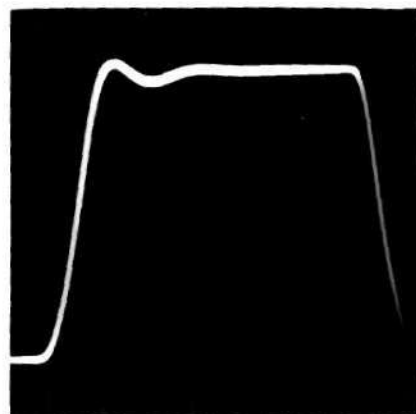


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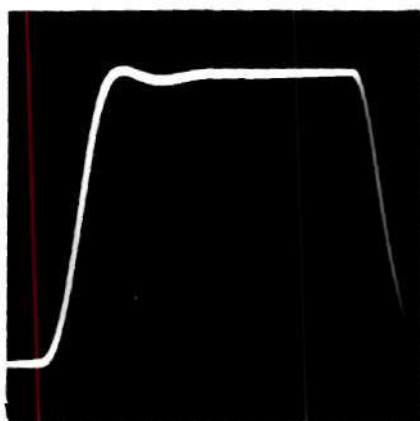


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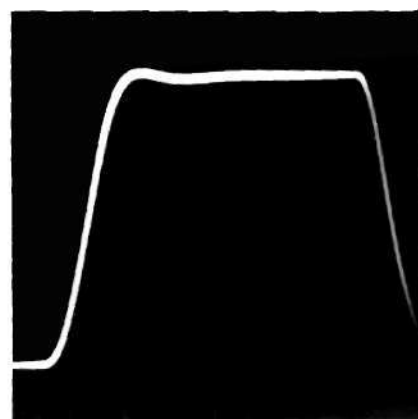


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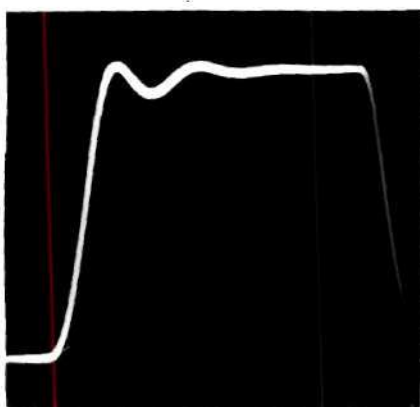


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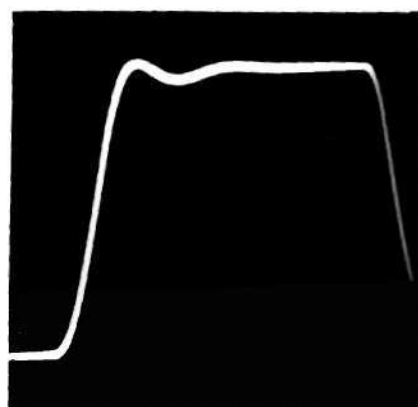


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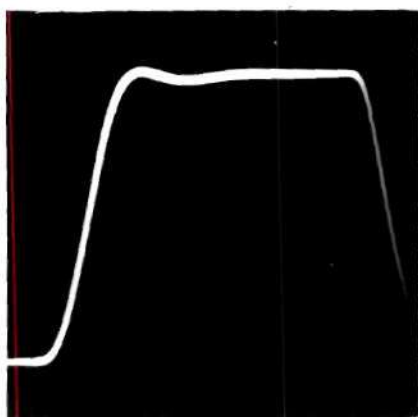


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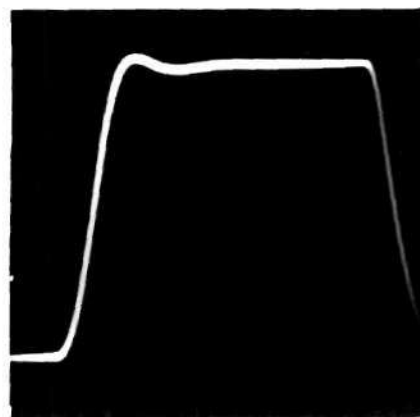


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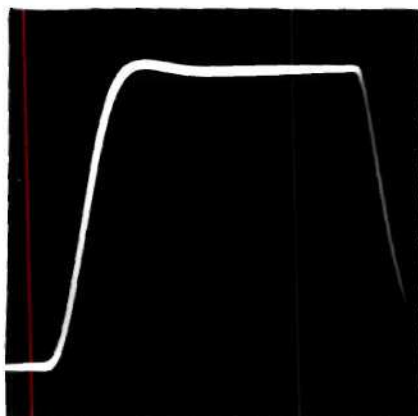


Fig. 110



Fig. 111

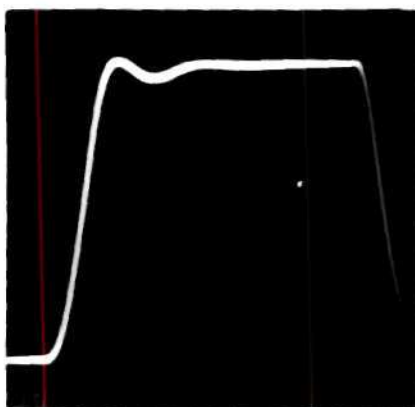


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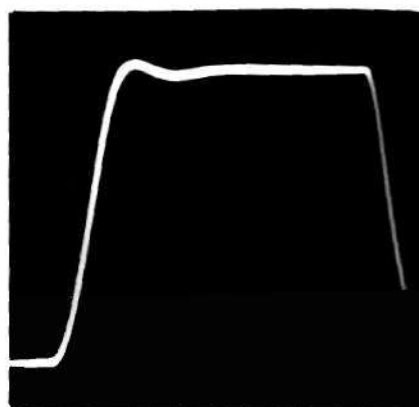


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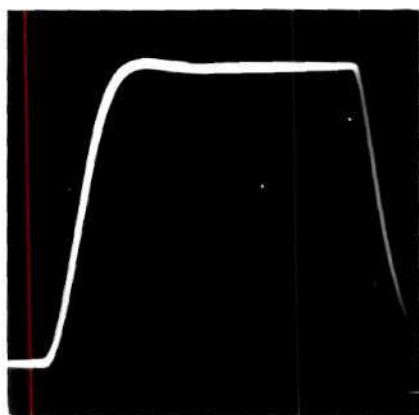


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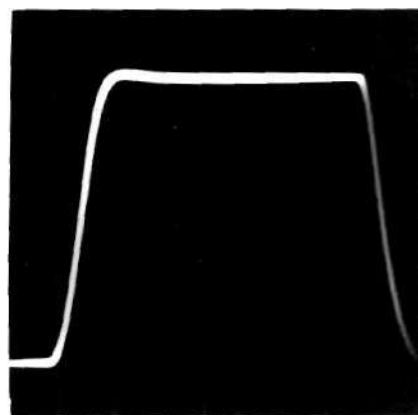


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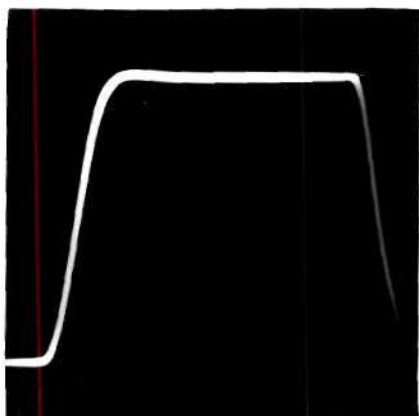


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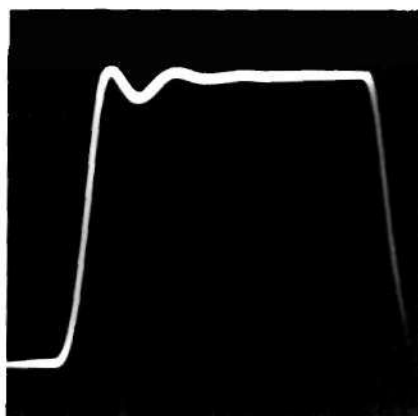


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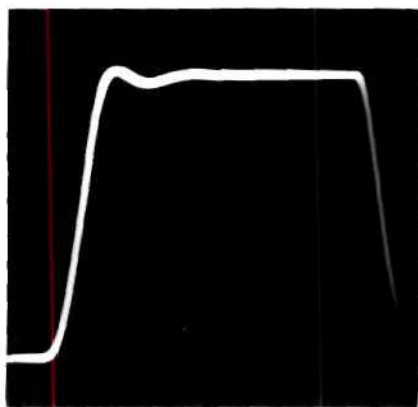


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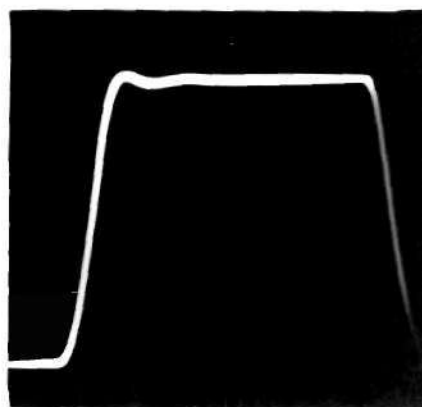


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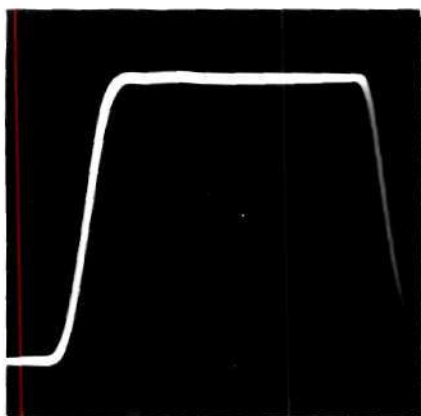


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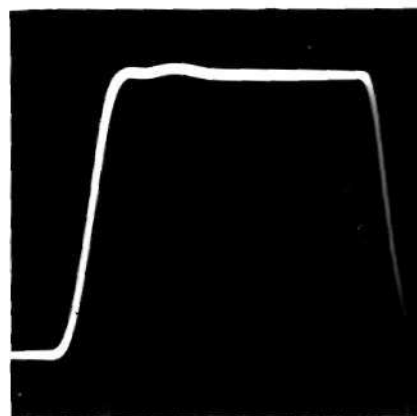


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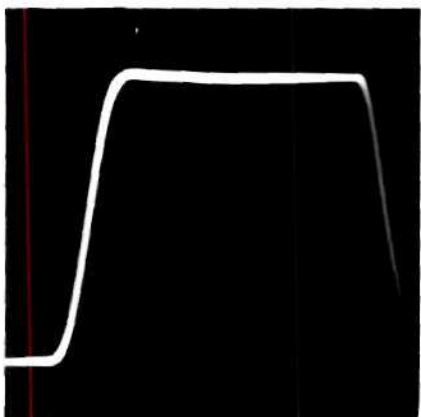


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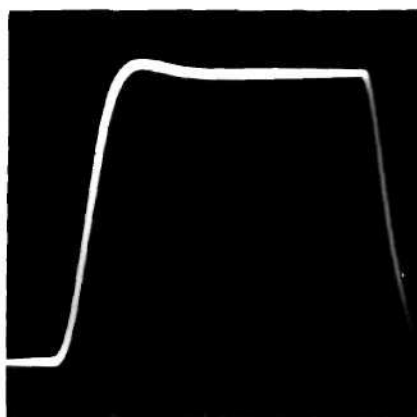


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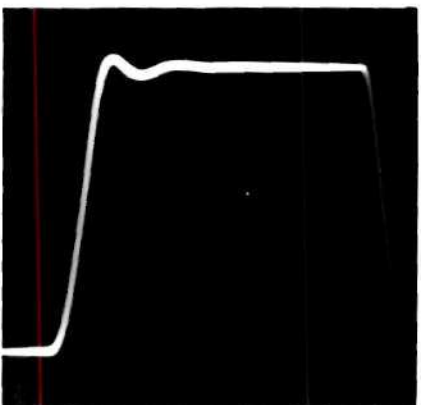


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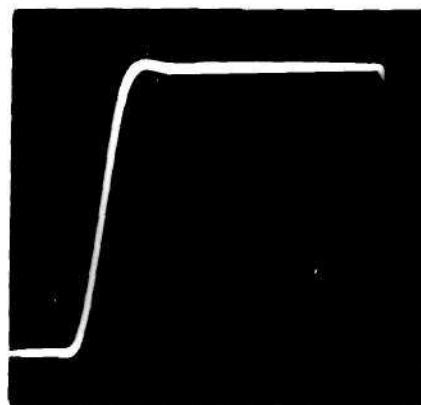


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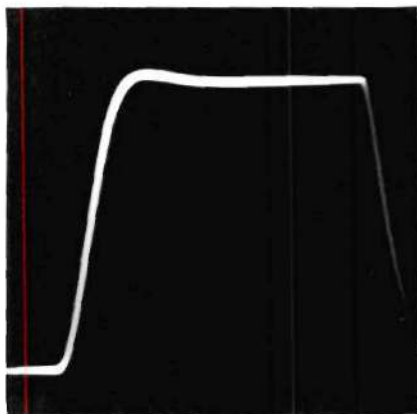


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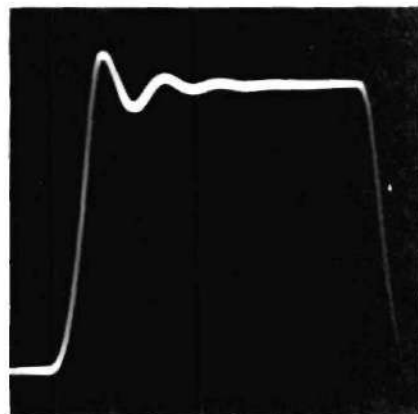


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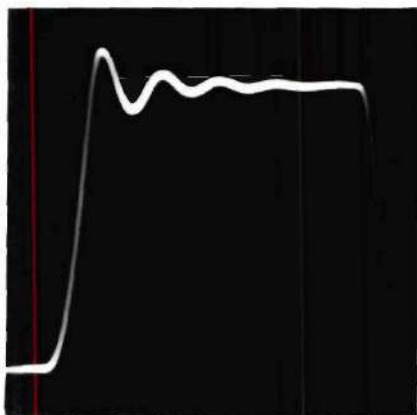


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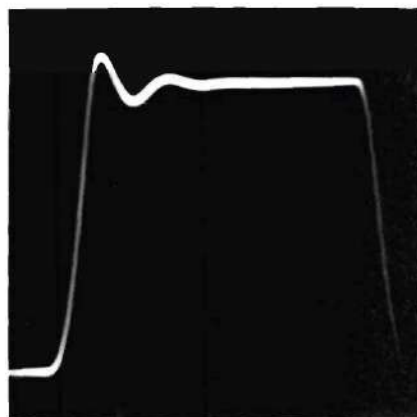


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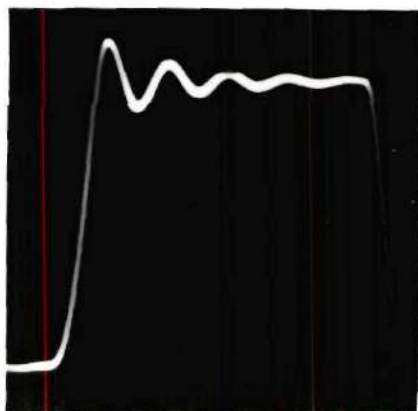


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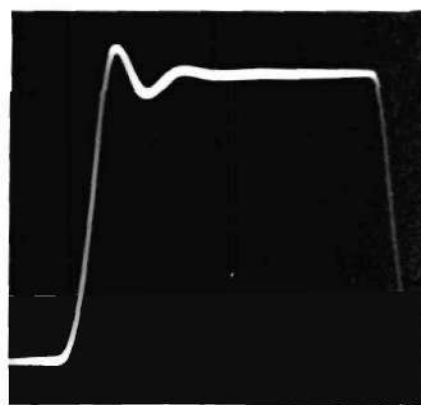


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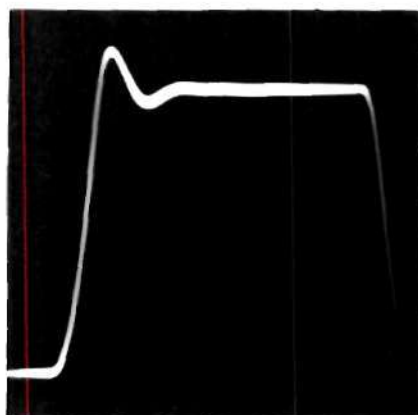


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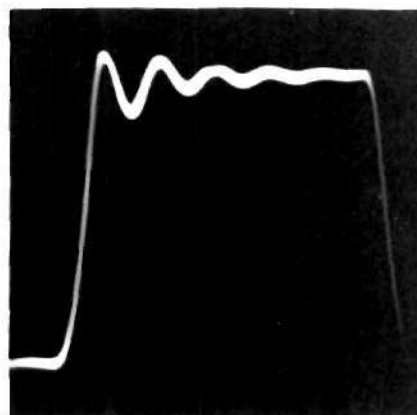


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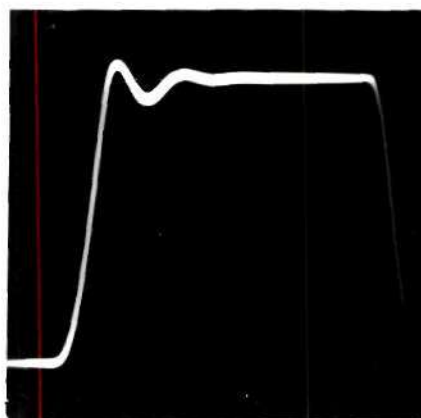


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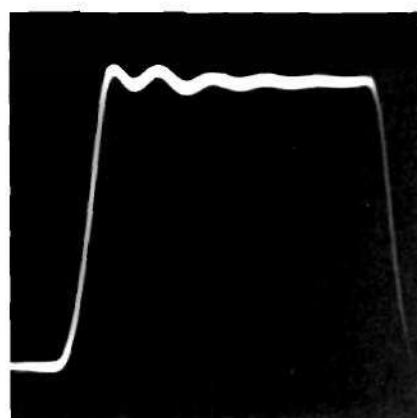


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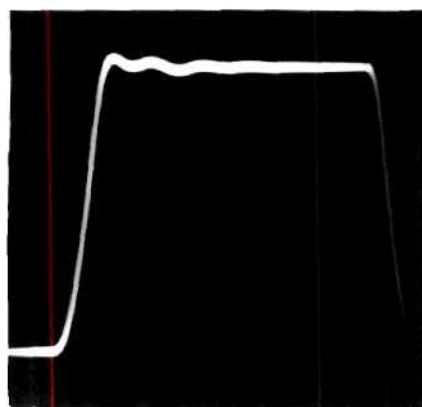


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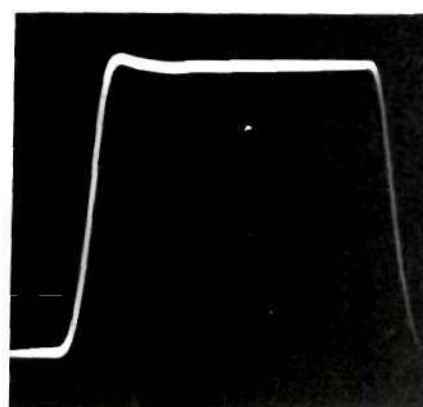


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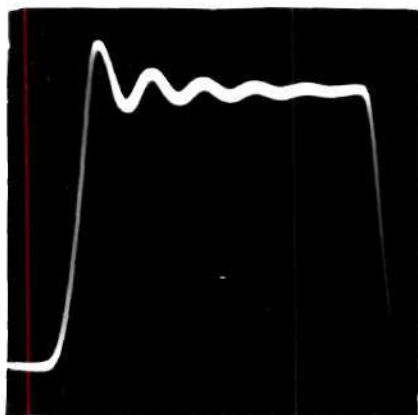


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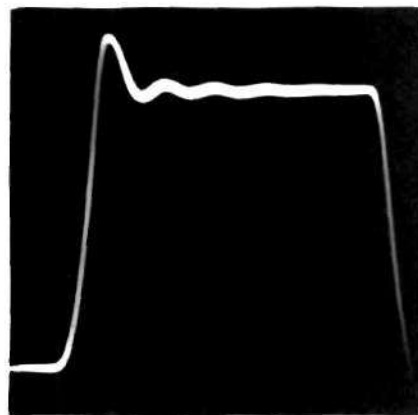


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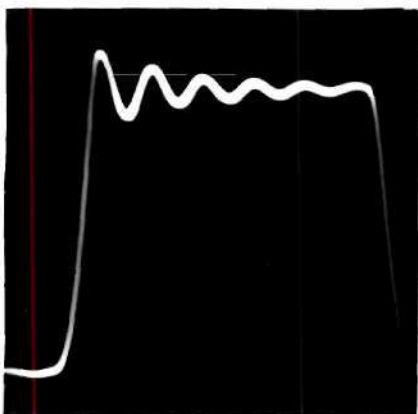


Fig. 140

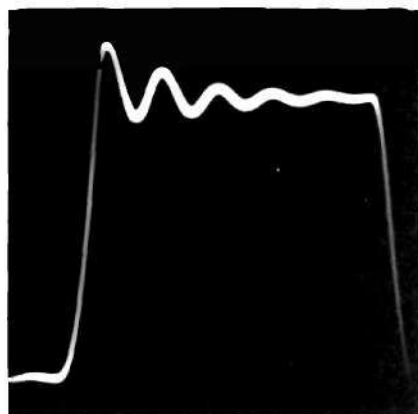


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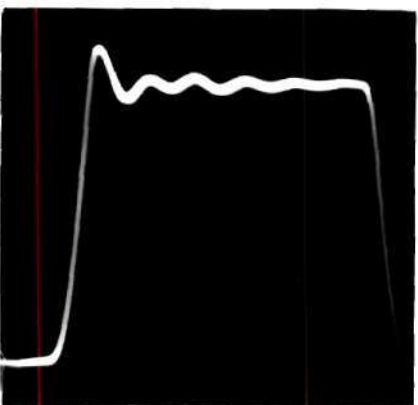


Fig. 142

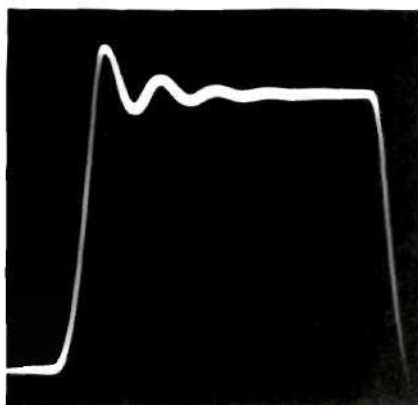


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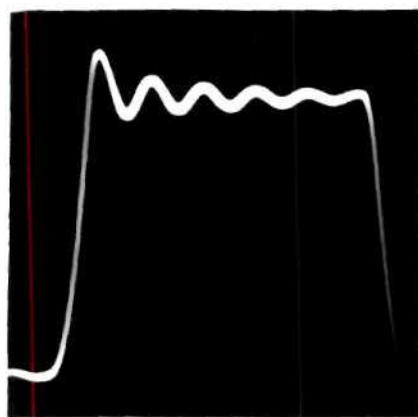


Fig. 144

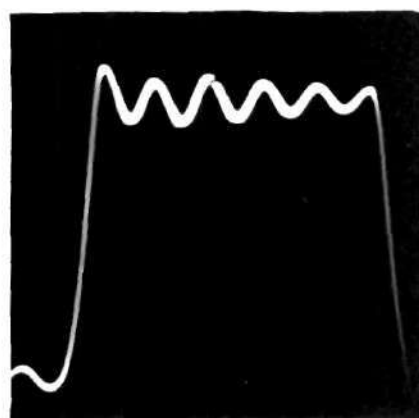


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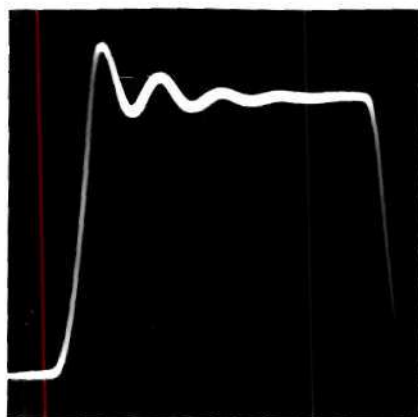


Fig. 146



Fig. 147

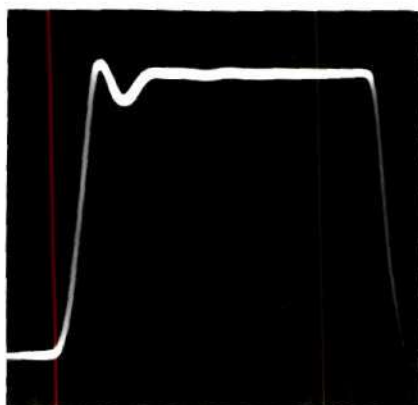


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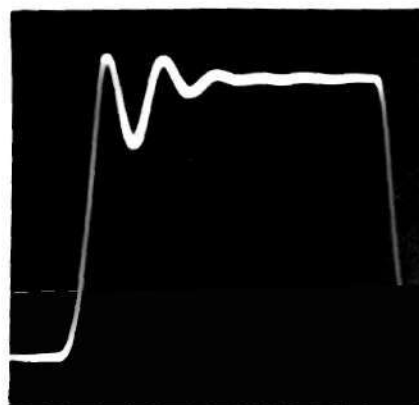


Fig. 149

CONCLUSION

Discussion of the Results

It is not the purpose of this thesis to represent a particular interstage network as being the most satisfactory one for all video work. The aim is to present information which will facilitate the design of a network to meet prescribed engineering requirements. Sufficient performance data are given to aid in the selection of a suitable configuration. A simple method of computing element values is described. It is left to the reader to draw conclusions regarding the merits of particular networks in which he may be interested, but it may be well to supplement the data with a few comments.

It is evident that the degree of improvement in performance afforded by the complex networks does not, in every case, justify their use. The selection of a configuration for a particular application depends upon the specifications to be met. It may easily be true that a very simple interstage will perform creditably enough to meet the requirements at hand. In fact, it does not follow that a complex network will always give better results than a simpler one.

Four-terminal networks were, as anticipated, capable of steeper rise reproduction than were their two-terminal

counterparts. However, the advantage of the former over the latter as a class was not striking in this respect.

The oscillatory tendencies of the networks were much more pronounced in the complex types. Readjustments of the element values were generally necessary to reduce these effects. In some cases it was not possible to reduce the oscillations to proportions which are tolerable in common applications.

The very simple compensation schemes which involved the use of small cathode by-pass condensers (Networks 2 and 3) gave excellent rise time performances but it should be pointed out that these circuits have a serious limitation. In the practical video amplifier, rather large cathode by-pass condensers are normally used. Any drastic reduction in the size of such a component results in a large loss of gain due to the introduction of negative feedback.⁵⁴ The amplifier stage which was used in the experiments was constructed with no cathode by-pass condenser at all, so no loss of gain resulted from the addition of a small one to the circuit, in fact the gain actually increased slightly when this change was made. The results which were obtained with this compensation system are, therefore, somewhat misleading.

The resistances of the series peaking coils in Networks

⁵⁴Sarbacher and Edson, op. cit., p. 463.

9 and 10 were included in the data because damping resistors were later added to these networks (Networks 13, 14, 15 and 16) and the amounts of damping initially present are of interest when evaluating the results.

Application of the Test Results to Interstage Design

This is the last topic to be discussed and should prove to be of practical interest. Given a set of specifications, the procedure to be followed in designing a video stage to amplify sharp pulses may be outlined as follows (high-frequency considerations only):

1. Consider first an uncompensated resistance-coupled stage. Determine C_t by any practical method.⁵⁵
2. Determine the time constant ($T = R_L C_t$) which the uncompensated stage must have to meet the rise time tolerances.
3. Compute R_L from the value of T obtained in step 2.
4. Calculate the gain obtainable from the given tube with the R_L determined in step 3. If it is not high enough to meet the requirements, consult the experimental data. Select a network which will give the required improvement. The value of R_L may now be increased to produce the desired gain and the rise time limit will not be exceeded because of the inherent advantage of the network configuration.

⁵⁵ See the article cited in footnote 8 for a method which utilizes a Q-meter.

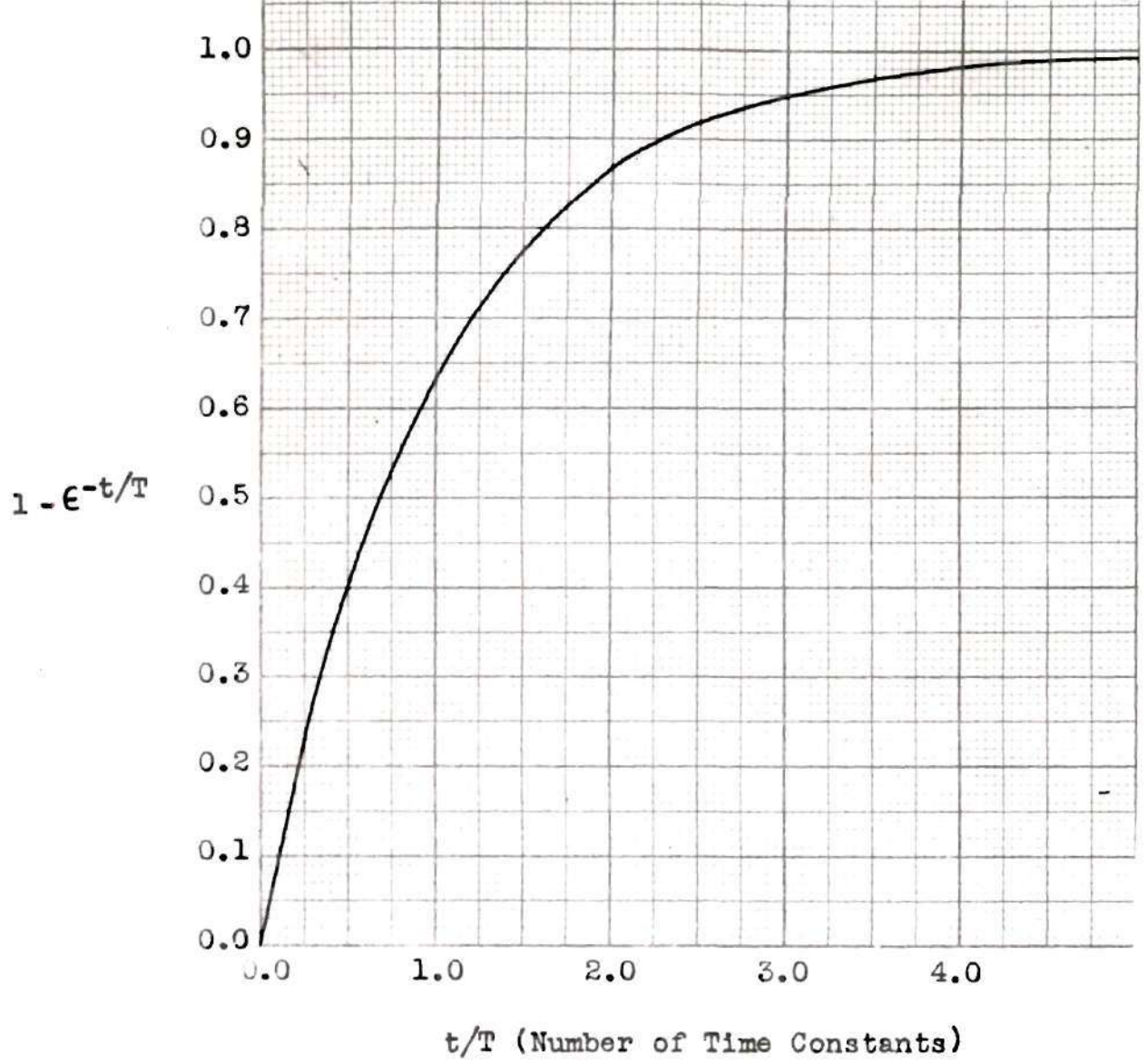
5. Increase R_L to the value dictated by the gain specification. Obtain the proper values of the other elements in the selected network by applying the reduction process described in Appendix I to the element values used in the test amplifier.

The method is best explained by an example. Suppose that the problem is to design a video stage capable of reproducing a rectangular pulse with an output rise time of not more than 0.1 microsecond. The pulse duration is of no concern here since it is related only to low-frequency characteristics. The tube available is a 6AC7 ($g_m = 9000$ micromhos) and the required voltage amplification is 45.

Before proceeding, the term "rise time" must be carefully defined. It is common practice to define the rise time as the time required for the pulse amplitude to increase from 10 per cent to 90 per cent of its final value. The transition takes place, in a simple RC circuit, in a manner defined by

$$e = E_{\max}(1 - e^{-t/T})$$

where $T = RC$ is the time constant of the circuit and e is the amplitude at the time t . Reference to Figure 150(a) will show that the time required for the pulse to rise to 10 per cent of its full height is about 0.1 of a time constant. The relative amplitude is 90 per cent after 2.3 time constants have elapsed. Hence, the rise time, by



(a) Graph of the Function $1 - e^{-t/T}$

(b) Measurement of Rise Time

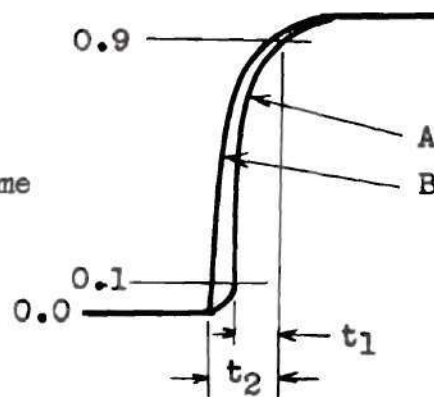


Fig. 150

definition, is

$$2.3t/T - 0.1t/T = 2.2t/T \text{ or } 2.2 \text{ time constants.}$$

The reason for measuring the rise time from 10 per cent amplitude rather than from zero is illustrated in Figure 150(b). A and B represent the hypothetical output waves of a four-terminal interstage and a two-terminal interstage respectively. It is seen that waveform A is shifted to the right and has a rounded corner at the bottom of the leading edge. The effect is exaggerated for clarity, but the point is that rise time measurements from zero amplitude as a reference would give a value of t_2 for the four-terminal network and a little less than that for the other interstage. The value of t_1 for the four-terminal network is evidently more meaningful.

The rise time measurement is made on the basis of 90 per cent relative height because this value is easy to locate graphically, and confusion due to overshoot is thereby evaded.

To proceed with the design problem in the sequence of steps previously outlined:

1. The amplifier is constructed with resistance-coupling and C_t is found to be 25 micro-microfarads. It is wired in such a manner that compensating elements may be inserted later without appreciably changing C_t .

$$2. \quad T = \frac{\text{rise time}}{2.2} = \frac{10^{-7} \text{ sec.}}{2.2} = 4.54 \times 10^{-8} \text{ second}$$

$$3. \quad R_L = T/C_t = \frac{4.54 \times 10^{-8}}{25 \times 10^{-12}} = 1816 \text{ ohms}$$

$$4. \quad \text{Gain} = g_m R_L = 9 \times 10^{-3} \times 1816 = 16.34$$

This is not adequate, so a compensating network must be selected. Figure 137 indicates that Network 17 is quite effective. A rise time measurement on the figure yields a length of 0.15 inch. A numerical advantage ratio may be obtained in relation to the performance of an uncompensated stage tested under the same conditions. Network 17 was tested with a 2500 cycle square wave, so Figure 13, the output oscillogram of an uncompensated stage tested under identical conditions, is checked for rise time. The length, in this case, is 0.45 inch. The rise time is, therefore, improved by a factor of three when Network 17 is used. Accordingly, T may be increased by a factor of three or R_L may be raised to 5448 ohms without exceeding the rise time tolerance.

5. To obtain a gain of 45,

$$R_L = \frac{45}{9 \times 10^{-3}} = 5000 \text{ ohms}$$

which is less than the maximum value permitted. As a

compromise, let $R_L = 5200$ ohms, this value being on the safe side with respect to both rise time and gain requirements. R_L was 1000 ohms in the test amplifier, so the first reduction factor is 5.2. The other values used in the experiment were:

$$C_p = 0.007 \text{ mfd.}, C_g = 0.013 \text{ mfd.}, C_3 = 0.004 \text{ mfd.}, \\ L_1 = 2.5 \text{ mh.}, L_2 = 11.0 \text{ mh.}$$

$$\begin{aligned} (b)^* C_p &= 7 \times 10^{-9} \times 1/5.2 = 1.35 \times 10^{-9} \text{ farad} \\ C_g &= 13 \times 10^{-9} \times 1/5.2 = 2.5 \times 10^{-9} \text{ farad} \\ C_3 &= 4 \times 10^{-9} \times 1/5.2 = 0.77 \times 10^{-9} \text{ farad} \\ L_1 &= 2.5 \times 10^{-3} \times 5.2 = 13.00 \times 10^{-3} \text{ henry} \\ L_2 &= 11.0 \times 10^{-3} \times 5.2 = 57.2 \times 10^{-3} \text{ henry} \end{aligned}$$

(c) The actual value of C_t will be 25×10^{-12} farad.

$$\text{Factor} = \frac{25 \times 10^{-12}}{3.85 \times 10^{-9}} = 6.49 \times 10^{-3}$$

(d) The final values are:

$$\begin{aligned} \underline{C_p} &= 1.35 \times 10^{-9} \times 6.49 \times 10^{-3} = 8.76 \times 10^{-12} \\ &\text{farad or } \underline{8.76 \text{ mmfd.}} \\ \underline{C_g} &= 2.5 \times 10^{-9} \times 6.49 \times 10^{-3} = 16.2 \times 10^{-12} \\ &\text{farad or } \underline{16.2 \text{ mmfd.}} \\ \underline{C_3} &= 0.77 \times 10^{-9} \times 6.49 \times 10^{-3} = 5.00 \times 10^{-12} \\ &\text{farad or } \underline{5.00 \text{ mmfd.}} \end{aligned}$$

*The steps are labeled to correspond with the discussion in Appendix I. Step (a) is included in 5.

$$\underline{L_1} = 13.00 \times 10^{-3} \times 6.49 \times 10^{-3} = 84.4 \times 10^{-6}$$

henry or 84.4 microhenries

$$\underline{L_2} = 57.2 \times 10^{-3} \times 6.49 \times 10^{-3} = 371 \times 10^{-6}$$

henry or 371 microhenries

It is felt that the method of design which has been presented reduces the problem to its simplest form. The results are, of course, subject to the errors introduced by inaccuracies in the techniques used to obtain the design factors. However, it is easily possible to produce designs which are within normal engineering tolerances.

The material presented in the preceding sections of the thesis should be of value in guiding the reader in a more detailed study of the video compensation problem, should he wish to pursue the subject further.

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APPENDIX I

Sample Calculations

Selection of C_t

Assume that the square wave frequency to be used is 2500 cycles. Each half-period is then $1/5000$ second in length. It was stated in the discussion of the experimental work that the square wave frequency was selected so that one half-period was ten time constants ($R_L C_t$) long. The load resistance, R_L , was 1000 ohms in all cases.

Therefore, $1/5000 = 10(1000C_t)$

$C_t = 1/5 \times 10^{-7} = 0.02 \times 10^{-6}$ farad or 0.02 microfarad.

Calculation of Values of Circuit Constants

It is conventional to present network designs in reduced or normalized form. Algebraic manipulations and numerical reductions are required to obtain the desired values in a particular case. Since several of the more complex networks presented were taken from Bode's work, a sample calculation using his notation is presented

Bode expresses his circuit element values on a per unit basis.⁵⁶ "The numbers refer to a filter with unit cut-off and unit impedance level." These values must be multiplied by suitable factors to raise the frequency and

⁵⁶Bode, op. cit., p. 429 (footnote).

impedance levels to meet existing conditions.

Computations may be carried out in the following sequence:

(a) Multiply the number given for R_L by the factor which will raise it to the proper level.

(b) Decrease the capacitances and increase the inductances by the factor used in (a). If there were other resistances in the network they would also be increased by this same factor.

(c) Multiply the C_t resulting from (b) by the factor which will reduce it to its actual value.

(d) Multiply all other capacitances and all inductances by the factor used in (c).

The procedure is equivalent to two operations:

1. Holding the frequency constant and raising the impedance level [steps (a) and (b)].
2. Holding the impedances constant and varying the frequency [steps (c) and (d)].

To illustrate, consider the following design numbers which are given for Network 7:

$$R_L = 1.0, C_t/2 = 1.0, L_1 = 2.00, L_2 = 0.600, C_3 = 1.600, \\ C_4 = 1.067$$

Let $f = 2500$ cycles and $R_L = 1000$ ohms. Then $C_t = 0.02$ mfd.

Step (a) $R_L = 1.0 \times 10^3 = 1000$ ohms. The factor is 10^3 .

Step (b) $C_t = 2.00 \times 10^{-3}$ farad

$$C_3 = 1.600 \times 10^{-3} \text{ farad}$$

$$C_4 = 1.067 \times 10^{-3} \text{ farad}$$

$$L_1 = 2.00 \times 10^3 \text{ henries}$$

$$L_2 = 0.600 \times 10^3 \text{ henries}$$

Step (c) $C_t = 2.00 \times 10^{-3} \times 10^{-5} = 2.0 \times 10^{-8}$ farad.

The factor is 10^{-5} .

Step (d) $C_3 = 1.600 \times 10^{-3} \times 10^{-5} = 1.6 \times 10^{-8}$

farad or 0.016 mfd.

$$C_4 = 1.067 \times 10^{-3} \times 10^{-5} = 1.067 \times 10^{-8}$$

farad or 0.01067 mfd.

$$L_1 = 2.00 \times 10^3 \times 10^{-5} = 2.00 \times 10^{-2}$$

henry or 20 mh.

$$L_2 = 0.600 \times 10^3 \times 10^{-5} = 0.600 \times 10^{-2}$$

henry or 6 mh.

APPENDIX II

Table of Measured Inductances and D-C Resistances
of Coils Used in Experiments

Inductance Box Number	Nominal Inductance, Mh.	Measured Inductance, Mh.	Measured D-C Resistance, Ohms

1.	0.50	0.49	5.16
	1.00	1.00	7.42
	2.00	2.01	13.92
	4.00	3.97	16.86
	8.00	8.00	24.30
2.	0.50	0.50	6.61
	1.00	1.00	13.56
	2.00	1.98	16.18
	4.00	3.90	23.68
	8.00	7.95	26.35
3.	0.50	0.50	4.27
	1.00	1.00	10.14
	2.00	1.98	36.93
	4.00	4.00	29.02
	8.00	7.95	29.92
4.	0.50	0.49	4.22
	1.00	1.00	6.72
	2.00	1.99	11.82
	4.00	3.95	20.91
	8.00	8.00	29.87